JC07 Rec'd PCT/PTO 04 JAN 2002

FORM PTO-1390 (Modified) (REV 11-2000) U.S. DEPARTMENT OF COMMERCE PATENT AND TRADEMARK OFFICE

TRANSMITTAL LETTER TO THE UNITED STATES DESIGNATED/ELECTED OFFICE (DO/EO/US) CONCERNING A FILING UNDER 35 U.S.C. 371

ATTORNEY'S DOCKET NUMBER
MTR.0031US

U.S. APPLICATION NO. (IF KNOWN, SEE 37 CFR
10/019961

INTERNATIONAL APPLICATION NO.	INTERNATIONAL FILING DATE	PRIORITY DATE CLAIMED
PCT/FR00/01904	**4 July 2000 (04.07.2000)**	**5 July 1999 (05.07.1999)**

TITLE OF INVENTION

METHODS AND DEVICE FOR AUDIO ANALYSIS AND SYNTHESIS

APPLICANT(S) FOR DO/EO/US

FRANCOIS CAPMAN and CARLO MURGIA

Applicant herewith submits to the United States Designated/Elected Office (DO/EO/US) the following items and other information:

1. ☒ This is a **FIRST** submission of items concerning a filing under 35 U.S.C. 371.
2. ☐ This is a **SECOND** or **SUBSEQUENT** submission of items concerning a filing under 35 U.S.C. 371.
3. ☒ This is an express request to begin national examination procedures (35 U.S.C. 371(f)). The submission must include itens (5), (6), (9) and (24) indicated below.
4. ☒ The US has been elected by the expiration of 19 months from the priority date (Article 31).
5. ☒ A copy of the International Application as filed (35 U.S.C. 371 (c) (2))
   a. ☒ is attached hereto (required only if not communicated by the International Bureau).
   b. ☐ has been communicated by the International Bureau.
   c. ☐ is not required, as the application was filed in the United States Receiving Office (RO/US).
6. ☒ An English language translation of the International Application as filed (35 U.S.C. 371(c)(2)).
   a. ☒ is attached hereto.
   b. ☐ has been previously submitted under 35 U.S.C. 154(d)(4).
7. ☐ Amendments to the claims of the International Application under PCT Article 19 (35 U.S.C. 371 (c)(3))
   a. ☐ are attached hereto (required only if not communicated by the International Bureau).
   b. ☐ have been communicated by the International Bureau.
   c. ☐ have not been made; however, the time limit for making such amendments has NOT expired.
   d. ☐ have not been made and will not be made.
8. ☐ An English language translation of the amendments to the claims under PCT Article 19 (35 U.S.C. 371(c)(3)).
9. ☒ An oath or declaration of the inventor(s) (35 U.S.C. 371 (c)(4)).
10. ☐ An English language translation of the annexes to the International Preliminary Examination Report under PCT Article 36 (35 U.S.C. 371 (c)(5)).
11. ☒ A copy of the International Preliminary Examination Report (PCT/IPEA/409).
12. ☐ A copy of the International Search Report (PCT/ISA/210).

Items 13 to 20 below concern document(s) or information included:

13. ☐ An Information Disclosure Statement under 37 CFR 1.97 and 1.98.
14. ☒ An assignment document for recording. A separate cover sheet in compliance with 37 CFR 3.28 and 3.31 is included.
15. ☒ A **FIRST** preliminary amendment.
16. ☐ A **SECOND** or **SUBSEQUENT** preliminary amendment.
17. ☐ A substitute specification.
18. ☐ A change of power of attorney and/or address letter.
19. ☐ A computer-readable form of the sequence listing in accordance with PCT Rule 13ter.2 and 35 U.S.C. 1.821 - 1.825.
20. ☐ A second copy of the published international application under 35 U.S.C. 154(d)(4).
21. ☐ A second copy of the English language translation of the international application under 35 U.S.C. 154(d)(4).
22. ☒ Certificate of Mailing by Express Mail
23. ☒ Other items or information:

   Eleven (11) sheets of formal drawings.

 PCTUS1/REV03

Express Mail No. EL857974173US

JC13 Rec'd PCT/PTO 04 JAN 2002

U.S. APPLICATION NO. (IF KNOWN, SEE 37 CFR	INTERNATIONAL APPLICATION NO.	ATTORNEY'S DOCKET NUMBER
10/019961	PCT/FR00/01904	MTR.0031US

24. The following fees are submitted:.

BASIC NATIONAL FEE (37 CFR 1.492 (a) (1) - (5)) :

		CALCULATIONS	PTO USE ONLY
☐ Neither international preliminary examination fee (37 CFR 1.482) nor international search fee (37 CFR 1.445(a)(2)) paid to USPTO and International Search Report not prepared by the EPO or JPO	**$1040.00**		
☒ International preliminary examination fee (37 CFR 1.482) not paid to USPTO but International Search Report prepared by the EPO or JPO	**$890.00**		
☐ International preliminary examination fee (37 CFR 1.482) not paid to USPTO but international search fee (37 CFR 1.445(a)(2)) paid to USPTO	**$740.00**		
☐ International preliminary examination fee (37 CFR 1.482) paid to USPTO but all claims did not satisfy provisions of PCT Article 33(1)-(4)	**$710.00**		
☐ International preliminary examination fee (37 CFR 1.482) paid to USPTO and all claims satisfied provisions of PCT Article 33(1)-(4)	**$100.00**		
ENTER APPROPRIATE BASIC FEE AMOUNT =		**$890.00**	
Surcharge of **$130.00** for furnishing the oath or declaration later than ☐ 20 ☐ 30 months from the earliest claimed priority date (37 CFR 1.492 (e)).		**$0.00**	

CLAIMS	NUMBER FILED	NUMBER EXTRA	RATE		
Total claims	28 - 20 =	8	x $18.00	**$144.00**	
Independent claims	6 - 3 =	3	x $84.00	**$252.00**	
Multiple Dependent Claims **(check if applicable).**			☐	**$0.00**	
TOTAL OF ABOVE CALCULATIONS =				**$1,286.00**	
☐ Applicant claims small entity status. See 37 CFR 1.27). The fees indicated above are reduced by 1/2.				**$0.00**	
SUBTOTAL =				**$1,286.00**	
Processing fee of **$130.00** for furnishing the English translation later than ☐ 20 ☐ 30 months from the earliest claimed priority date (37 CFR 1.492 (f)). +				**$0.00**	
TOTAL NATIONAL FEE =				**$1,286.00**	
Fee for recording the enclosed assignment (37 CFR 1.21(h)). The assignment must be accompanied by an appropriate cover sheet (37 CFR 3.28, 3.31) **(check if applicable).**			☒	**$40.00**	
TOTAL FEES ENCLOSED =				**$1,326.00**	
				Amount to be: refunded	$
				charged	$

a. ☒ A check in the amount of **$1,326.00** to cover the above fees is enclosed.

b. ☐ Please charge my Deposit Account No. ________ in the amount of ________ to cover the above fees. A duplicate copy of this sheet is enclosed.

c. ☒ The Commissioner is hereby authorized to charge any additional fees which may be required, or credit any overpayment to Deposit Account No. **20-1504** A duplicate copy of this sheet is enclosed.

d. ☐ Fees are to be charged to a credit card. **WARNING:** Information on this form may become public. **Credit card information should not be included on this form.** Provide credit card information and authorization on PTO-2038.

NOTE: Where an appropriate time limit under 37 CFR 1.494 or 1.495 has not been met, a petition to revive (37 CFR 1.137(a) or (b)) must be filed and granted to restore the application to pending status.

SEND ALL CORRESPONDENCE TO:

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21906
PATENT TRADEMARK OFFICE

SIGNATURE

Dan C. Hu
NAME

40,025
REGISTRATION NUMBER

January 4, 2002
DATE

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

Applicants:	Francois Capman et al.	Group Art Unit:	
Int'l Appl. No.:	PCT/FR00/01904	Examiner:	
Int'l Filing Date:	4 July 2000		
For:	Methods and Device for Audio Analysis and Synthesis	Atty. Dkt. No.:	MTR.0031US

Box PCT
Commissioner for Patents
Washington DC 20231

PRELIMINARY AMENDMENT

Sir:

Prior to Examination, please amend the above-identified application as follows

In the Specification:

Page 1, at line 2, please insert the following paragraph:

--BACKGROUND OF THE INVENTION--

Page 2, at line 30, please insert the following paragraph:

--SUMMARY OF THE INVENTION--

Page 5, delete lines 5-8.

Page 5, at line 9, please insert the following paragraph:

--BRIEF DESCRIPTION OF THE DRAWINGS--

Page 6, at line 17, insert the following paragraph:

--DETAILED DESCRIPTION--

In the Abstract:

--The N samples of a frame of an audio signal are weighted by an analysis window of Hamming, Hanning, Kaiser or similar type A spectrum of the audio signal is calculated by transforming each frame of weighted samples in the frequency domain, and the spectrum of the audio signal is processed to deliver parameters for synthesizing a signal derived from the

analyzed audio signal. The successive frames comprise an alternation of frames for which are delivered complete sets of synthesis parameters and of frames for which are delivered incomplete sets of synthesis parameters. The successive frames for which complete sets of synthesis parameters are delivered exhibit mutual overlaps of less than N/2 samples.--

<u>In the Claims:</u>

Amend the following claims:

1. (Amended) A method of analyzing an audio signal processed by successive frames of N samples, N being an integer greater than 1, comprising the steps of:

weighting the samples of each frame by an analysis window of Hamming, Hanning, Kaiser or similar type;

calculating a spectrum of the audio signal by transforming each frame of weighted samples in the frequency domain; and

processing the spectrum of the audio signal to deliver synthesis parameters for a signal derived from the analyzed audio signal;

wherein the successive frames comprise an alternation of frames for which complete sets of synthesis parameters are delivered and of frames for which incomplete sets of synthesis parameters are delivered, and wherein the successive frames for which complete sets of synthesis parameters are delivered exhibit mutual overlaps of less than N/2 samples.

2. (Amended) The method as claimed in claim 1, wherein the incomplete sets of synthesis parameters include data representing an error of interpolation of at least one of the synthesis parameters.

3. (Amended) The method as claimed in claim 1, wherein the incomplete sets of synthesis parameters include data representing a filter for interpolating at least one of the synthesis parameters.

4. (Amended) The method as claimed in claim 1, wherein the processing of the spectrum of the audio signal comprises extracting coding parameters for transmitting and/or storing a coded audio signal.

5. (Amended) The method as claimed in claim 1, wherein the processing of the spectrum of the audio signal comprises a denoising operation by spectral subtraction.

6. (Amended) An audio processing device, for analyzing an audio signal by successive frames of N samples, N being an integer greater than 1, comprising:

means for weighting the samples of each frame by an analysis window of Hamming, Hanning, Kaiser or similar type;

means for calculating a spectrum of the audio signal by transforming each frame of weighted samples in the frequency domain; and

means for processing the spectrum of the audio signal to deliver synthesis parameters for a signal derived from the analyzed audio signal;

wherein the successive frames comprise an alternation of frames for which complete sets of synthesis parameters are delivered and of frames for which incomplete sets of synthesis parameters are delivered, and wherein the successive frames for which complete sets of synthesis parameters are delivered exhibit mutual overlaps of less than N/2 samples.

7. (Amended) The device as claimed in claim 6, wherein the incomplete sets of synthesis parameters include data representing an error of interpolation of at least one of the synthesis parameters.

8. (Amended) The device as claimed in claim 6, wherein the incomplete sets of synthesis parameters include data representing a filter for interpolating at least one of the synthesis parameters.

9. (Amended) The device as claimed in claim 6, wherein the processing means comprise means for extracting coding parameters for transmitting and/or storing a coded audio signal.

10. (Amended) The device as claimed in claim 6, wherein the processing means comprise spectral subtraction means for cancelling noise in the audio signal.

11. (Amended) A method of synthesizing an audio signal, comprising the steps of:

obtaining successive spectral estimates respectively corresponding to frames of N samples of the audio signal weighted by an analysis window, N being an integer greater than 1;

evaluating each frame of the audio signal by transforming the spectral estimates in the time domain;

modifying each evaluated frame by applying thereto a processing corresponding to a division by said analysis window and to a multiplication by a synthesis window; and

synthesizing the audio signal as an overlap sum of the modified frames,

wherein the successive frames exhibit mutual overlaps of L samples, L being an integer greater than 1 and smaller than N/2,

and wherein the synthesis window $f_S(i)$ satisfies $f_S(N-L+i) + f_S(i) = A$ for $0 \leq i < L$, and $f_S(i) = A$ for $L \leq i < N-L$, A being a positive constant and i being a sample rank in a frame with $0 \leq i < N$.

12. (Amended) The method as claimed in claim 11, wherein the synthesis window $f_S(i)$ increases from 0 to A for i ranging from 0 to L.

13. (Amended) The method as claimed in claim 12, wherein the synthesis window $f_S(i)$ for $0 \leq i < L$ is a raised half-sinusoid.

14. (Amended) An audio processing device, comprising:

means for obtaining successive spectral estimates respectively corresponding to frames of N samples of an audio signal weighted by an analysis window, N being an integer greater than 1;

means for evaluating each frame of the audio signal by transforming the spectral estimates in the time domain;

means for modifying each evaluated frame by applying thereto a processing corresponding to a division by said analysis window and to a multiplication by a synthesis window; and

means for synthesizing the audio signal as an overlap sum of the modified frames,

wherein the successive frames exhibit mutual overlaps of L samples, L being an integer greater than 1 and smaller than N/2,

and wherein the synthesis window $f_S(i)$ satisfies $f_S(N-L+i) + f_S(i) = A$ for $0 \leq i < L$, and $f_S(i) = A$ for $L \leq i < N-L$, A being a positive constant and i being a sample rank in a frame with $0 \leq i < N$.

15. (Amended) The device as claimed in claim 14, wherein the synthesis window $f_S(i)$ increases from 0 to A for i ranging from 0 to L.

16. (Amended) The device as claimed in claim 15, wherein the synthesis window $f_S(i)$ for $0 \leq i < L$ is a raised half-sinusoid.

Add the following claims:

17. (New) A method of synthesizing an audio signal, comprising the steps of:

defining a set of successive overlapping frames of N samples of the audio signal, N being an integer greater than 1;

obtaining spectral estimates for a subset of the frames by processing synthesis parameters respectively associated with the frames of said subset;

obtaining spectral estimates for the frames of the set which are not in said subset, with an interpolation of at least part of the synthesis parameters;

evaluating the frames of the set weighted by an analysis window, by transforming in the time domain the spectral estimates respectively obtained for said frames; and

modifying each evaluated frame by applying thereto a processing corresponding to a division by said analysis window and to a multiplication by a synthesis window; and

synthesizing the audio signal as an overlap sum of the modified frames,

wherein the successive frames of said subset exhibit mutual time shifts of M samples, M being an integer greater than N/2, while the successive frames of said set exhibit mutual time shifts of M/p samples, p being an integer larger than 1,

and wherein, the samples of a frame having ranks i numbered from 0 to N-1, the synthesis window $f'_S(i)$ has a support limited to the ranks i ranging from N/2-M/p to N/2+M/p and satisfies $f'_S(i) + f'_S(i+M/p) = A$ for $N/2-M/p \leq i < N/2$, A being a positive constant.

18. (New) The method as claimed in claim 17, wherein the synthesis window $f'_S(i)$ increases for i ranging from N/2-M/p to N/2.

19. (New) The method as claimed in claim 18, wherein the synthesis window $f'_S(i)$ is a raised sinusoid for $N/2-M/p \leq i < N/2+M/p$.

20. (New) The method as claimed in claim 17, further comprising the steps of:

associating data representing an interpolation error with the frames which are not in said subset; and

correcting at least one of the interpolated synthesis parameters by means of said data.

21. (New) The method as claimed in claim 17, further comprising the steps of:

associating data representing an interpolator filter with the frames which are not in said subset; and

interpolating at least one of the synthesis parameters by means of the interpolator filter represented by said data.

22. (New) The method as claimed in claim 17, wherein the synthesis parameters comprise cepstral coefficients subjected to the interpolation.

23. (New) An audio processing device, comprising:

framing means for defining a set of successive overlapping frames of N samples of an audio signal, N being an integer greater than 1;

means for obtaining spectral estimates for a subset of the frames by processing synthesis parameters respectively associated with the frames of said subset;

means for obtaining spectral estimates for the frames of the set which are not in said subset, with an interpolation of at least part of the synthesis parameters;

means for evaluating the frames of the set weighted by an analysis window, by transforming in the time domain the spectral estimates respectively obtained for said frames; and

means for modifying each evaluated frame by applying thereto a processing corresponding to a division by said analysis window and to a multiplication by a synthesis window; and

means for synthesizing the audio signal as an overlap sum of the modified frames,

wherein the successive frames of said subset exhibit mutual time shifts of M samples, M being an integer greater than N/2, while the successive frames of said set exhibit mutual time shifts of M/p samples, p being an integer larger than 1,

and wherein, the samples of a frame having ranks i numbered from 0 to N-1, the synthesis window $f'_S(i)$ has a support limited to the ranks i ranging from N/2-M/p to N/2+M/p and satisfies $f'_S(i) + f'_S(i+M/p) = A$ for $N/2-M/p \leq i < N/2$, A being a positive constant.

24. (New) The device as claimed in claim 23, wherein the synthesis window $f'_S(i)$ increases for i ranging from N/2-M/p to N/2.

25. (New) The device as claimed in claim 24, wherein the synthesis window $f'_S(i)$ is a raised sinusoid for $N/2-M/p \leq i < N/2+M/p$.

26. (New) The device as claimed in claim 23, further comprising:

means for associating data representing an interpolation error with the frames which are not in said subset; and

means for correcting at least one of the interpolated synthesis parameters by means of said data.

27. (New) The device as claimed in claim 23, further comprising:

means for associating data representing an interpolator filter with the frames which are not in said subset; and

means for interpolating at least one of the synthesis parameters by means of the interpolator filter represented by said data.

28. (New) The device as claimed in claim 23, wherein the synthesis parameters comprise cepstral coefficients subjected to the interpolation.

Remarks:

Allowance of all claims is respectfully requested. The Commissioner is authorized to charge any additional fees under 37 C.F.R. § 1.16 and § 1.17, or credit any overpayment to Deposit Account No. 20-1504 (MTR.0030US).

Date: 1/4/02

Respectfully submitted,

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VERSIONS WITH MARKINGS TO SHOW CHANGES

IN THE CLAIMS:

New claims 17-28 have been added. Amendments of the claims are indicated below:

1. (Amended) A method of analyzing an audio signal [(x)] processed by successive frames of N samples, N being an integer greater than 1, comprising the steps of:

[in which] weighting the samples of each frame [are weighted] by an analysis window [(f_A)] of Hamming, Hanning, Kaiser or similar type;[,]

calculating a spectrum of the audio signal [is calculated] by transforming each frame of weighted samples in the frequency domain; and [,]

processing the spectrum of the audio signal [is processed so as] to deliver synthesis parameters [(cx_sup, cx_inf,Emix)] for [synthesizing] a signal derived from the analyzed audio signal;[,]

[characterized in that] wherein the successive frames comprise an alternation of frames for which [are delivered] complete sets of synthesis parameters are delivered and of frames for which [are delivered] incomplete sets of synthesis [parameters, and in that] parameters are delivered, and wherein the successive frames for which complete sets of synthesis parameters are delivered exhibit mutual overlaps of less than N/2 samples.

2. (Amended) The method as claimed in claim 1, [in which] wherein the incomplete sets of synthesis parameters include data [(icx[n-1/2)] representing an error [(ecx[n-1/2])] of interpolation of at least one of the synthesis parameters.

3. (Amended) The method as claimed in claim 1, [in which] wherein the incomplete sets of synthesis parameters include data [(iP)] representing a filter [(128)] for interpolating at least one of the synthesis parameters.

4. (Amended) The method as claimed in [in any one of claims 1 to 3, in which] claim 1, wherein the processing of the spectrum of the audio signal [(x)] comprises [an extraction of] extracting coding parameters [(cx_sup, cx_inf, Emix) with a view to the transmission and/or the store of the] for transmitting and/or storing a coded audio signal.

5. (Amended) The method as claimed in [any one of claims 1 to 3, in which] claim 1, wherein the processing of the spectrum of the audio signal [(x)] comprises a denoising operation by spectral subtraction.

6. (Amended) An audio processing device, [comprising analysis means for executing a method as claimed in claims 1 to 5] for analyzing an audio signal by successive frames of N samples, N being an integer greater than 1, comprising:

means for weighting the samples of each frame by an analysis window of Hamming, Hanning, Kaiser or similar type;

means for calculating a spectrum of the audio signal by transforming each frame of weighted samples in the frequency domain; and

means for processing the spectrum of the audio signal to deliver synthesis parameters for a signal derived from the analyzed audio signal;

wherein the successive frames comprise an alternation of frames for which complete sets of synthesis parameters are delivered and of frames for which incomplete sets of synthesis parameters are delivered, and wherein the successive frames for which complete sets of synthesis parameters are delivered exhibit mutual overlaps of less than N/2 samples.

7. (Amended) [A method of synthesizing an audio signal, in which successive spectral estimates (Y) corresponding respectively to frames of N samples of the audio signal which are weighted by an analysis window (f_A) are obtained, the successive frames exhibiting mutual overlaps of L samples, each frame of the audio signal is evaluated by transforming the spectral estimates in the time domain, and the frames evaluated are combined to form the synthesized signal ($\hat{x}$), characterized in that each evaluated frame is modified by applying thereto a processing corresponding to a division by said analysis window (f_A) and to a multiplication by a synthesis window (f_S), and the synthesized signal is formed as an overlap sum of the modified frames, and in that, the number L being smaller than N/2 and the samples of a frame having ranks i numbered from 0 to N-1, the synthesis window $f_S(i)$ satisfies $f_S(N-L+i) + f_S(i) = A$ for $0 \leq i < L$, and is equal to

A for $L \leq i < N-L$, A being a positive constant]The device as claimed in claim 6, wherein the incomplete sets of synthesis parameters include data representing an error of interpolation of at least one of the synthesis parameters.

8. (Amended) [The method as claimed in claim 7, in which the synthesis window $f_S(i)$ increases from 0 to A for i going from 0 to L] The device as claimed in claim 6, wherein the incomplete sets of synthesis parameters include data representing a filter for interpolating at least one of the synthesis parameters.

9. (Amended) [The method as claimed in claim 8, in which the synthesis window $f_S(i)$ for $0 \leq i < L$ is a raised half-sinusoid] The device as claimed in claim 6, wherein the processing means comprise means for extracting coding parameters for transmitting and/or storing a coded audio signal.

10. (Amended) [A method of synthesizing an audio signal, in which a set of successive overlapping frames of N samples of the audio signal which are weighted by an analysis window (f_A) is evaluated, by transforming in the time domain spectral estimates (Y) corresponding respectively to said frames, and the evaluated frames are combined to form the synthesized signal ($\hat{x}$), characterized in that, for a subset of the evaluated frames, the spectral estimates are obtained by processing synthesis parameters (cx_sup_q, cx_inf_q, Emix) respectively associated with the frames of said subset while, for the frames which do not form part of the subset, the spectral estimates are obtained with an interpolation of a part at least of the synthesis parameters, in that the successive frames of said subset exhibit mutual time shifts of M samples, the number M being larger than N/2, while the successive frames of said set exhibit mutual time shifts of M/p samples, p being an integer larger than 1, in that each evaluated frame is modified by applying thereto a processing corresponding to a division by said analysis window (f_A) and to a multiplication by a synthesis window (f'_S), and the synthesized signal is formed as an overlap sum of the modified frames, and in that, the samples of a frame having ranks i numbered from 0 to N-1, the synthesis window $f'_S(i)$ has a support limited to the ranks i ranging from N/2 - M/p to N/2 + M/p and satisfies $f'_S(i) + f'_S(i + M/p) = A$ for $N/2 - M/p \leq i < N/2$, A being a positive constant] The

device as claimed in claim 6, wherein the processing means comprise spectral subtraction means for cancelling noise in the audio signal.

11. (Amended) [The method as claimed in claim 10, in which] A method of synthesizing an audio signal, comprising the steps of:

obtaining successive spectral estimates respectively corresponding to frames of N samples of the audio signal weighted by an analysis window, N being an integer greater than 1;

evaluating each frame of the audio signal by transforming the spectral estimates in the time domain;

modifying each evaluated frame by applying thereto a processing corresponding to a division by said analysis window and to a multiplication by a synthesis window; and

synthesizing the audio signal as an overlap sum of the modified frames,

wherein the successive frames exhibit mutual overlaps of L samples, L being an integer greater than 1 and smaller than N/2,

and wherein the synthesis window [$f'_S(i)$ increases for i ranging from N/2-M/p to N/2] $f_S(i)$ satisfies $f_S(N-L+i) + f_S(i) = A$ for $0 \leq i < L$, and $f_S(i) = A$ for $L \leq i < N-L$, A being a positive constant and i being a sample rank in a frame with $0 \leq i < N$.

12. (Amended) The method as claimed in claim 11, [in which] wherein the synthesis window [$f'_S(I)$ for $N/2-M/p \leq i < N/2+M/p$ is a raised sinusoid] $f_S(i)$ increases from 0 to A for i ranging from 0 to L.

13. (Amended) The method as claimed in claim 12, wherein the synthesis window $f_S(i)$ for $0 \leq i < L$ is a raised half-sinusoid [The method as claimed in any one of claims 10 to 12, in which data (icx_q[n-1/2]) representing an interpolation error (ecx_q[n-1/2]) are associated with the frames which do not form part of said subset, and are used to correct at least one of the interpolated synthesis parameters (cx_i[n-1/2])].

14. (Amended) [The method as claimed in any one of claims 10 to 12, in which data (iP) representing an interpolator filter (128) are associated with the frames which do not form part of said subset, and are used to interpolate at least one of the synthesis parameters] An audio processing device, comprising:

means for obtaining successive spectral estimates respectively corresponding to frames of N samples of an audio signal weighted by an analysis window, N being an integer greater than 1;

means for evaluating each frame of the audio signal by transforming the spectral estimates in the time domain;

means for modifying each evaluated frame by applying thereto a processing corresponding to a division by said analysis window and to a multiplication by a synthesis window; and

means for synthesizing the audio signal as an overlap sum of the modified frames,

wherein the successive frames exhibit mutual overlaps of L samples, L being an integer greater than 1 and smaller than N/2,

and wherein the synthesis window $f_S(i)$ satisfies $f_S(N-L+i) + f_S(i) = A$ for $0 \leq i < L$, and $f_S(i) = A$ for $L \leq i < N-L$, A being a positive constant and i being a sample rank in a frame with $0 \leq i < N$.

15. (Amended) The device as claimed in claim 14, wherein the synthesis window $f_S(i)$ increases from 0 to A for i ranging from 0 to L [The method as claimed in any one of claims 10 to 14, in which the synthesis parameters comprise cepstral coefficients (cx[n]) subjected to the interpolation].

16. (Amended) [An audio processing device, comprising synthesis means for executing a method as claimed in any one of claims 7 to 15] The device as claimed in claim 15, wherein the synthesis window $f_S(i)$ for $0 \leq i < L$ is a raised half-sinusoid.

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METHODS AND DEVICES FOR AUDIO ANALYSIS AND SYNTHESIS

The present invention relates to the analysis and synthesis of audio signals, on the basis of representations of these signals in spectral domain.

It applies in particular, but not exclusively, to the coding of speech, in narrowband or in broadband, in various coding bit rate ranges. Among the other fields of application, mention may be made of denoising by spectral subtraction (see EP-A 0 534 837 or WO99/14739).

In the methods of analysis in question, the spectrum of the signal is obtained by transforming successive frames to the frequency domain. The transformation employed is usually the fast Fourier transform (FFT); however other known transforms can be used. In the frequent case of a sampling of the signal at 8 kHz, the number N of samples per frame is typically of the order of 100 to 500, this representing frames of a few tens of milliseconds. To benefit from the maximum resolution in frequency, the FFT is performed on 2N points, N zero samples being added to the N samples of the frame.

The spectrum obtained by Fourier transform of the signal frame is the convolution of the real spectrum of the signal by the Fourier transform of the signal analysis window. This analysis window, which weights the samples of each frame, is required so as to take account of the finite duration of the frame. If the signal frame is subjected to the FFT directly, that is to say if a rectangular analysis window is used, the spectrum obtained is disturbed by the secondary peaks of the FFT of the analysis window. To limit this drawback, which is especially noticeable when parameters representing the signal or the noise have to be extracted from the spectra, recourse is had to windows having better spectral properties, that is to

say weighting functions whose support is limited to N samples and whose Fourier transform has its energy concentrated in a narrow peak with a strong attenuation of the secondary peaks. The most common of these windows are the Hamming, Hanning and Kaiser windows.

In the analysis and synthesis procedure known as OLA ("Overlap-And-Add"), the successive frames exhibit mutual overlaps of 50% (N/2 samples). Since the analysis windows commonly used satisfy the property $f_A(i+N/2) + f_A(i) = 1$, synthesis can be performed simply by overlap-summing the frames of N samples, which frames are calculated in succession by inverse Fourier transform of the spectra.

With the aim of refining the spectral representation, certain procedures referred to as WOLA ("Weighted OLA") use, for analysis, frames whose mutual overlaps are more than 50%. For the synthesis, it is necessary to reweight the samples of the frames before summing them. These procedures increase the complexity of the analysis and of the synthesis. In coding applications, they also increase the transmission bit rate required.

An aim of the invention is to propose a scheme for analyzing and synthesizing audio signals which makes it possible to limit the rate of the analysis frames, while using analysis windows having good spectral properties.

The invention proposes a method of analyzing an audio signal processed by successive frames of N samples, in which the samples of each frame are weighted by an analysis window of Hamming, Hanning, Kaiser or similar type, a spectrum of the audio signal is calculated by transforming each frame of weighted samples in the frequency domain, and the spectrum of the audio signal is processed so as to deliver parameters for synthesizing a signal derived from the analyzed audio

signal. According to the invention, the successive frames comprise an alternation of frames for which are delivered complete sets of synthesis parameters, which exhibit mutual overlaps of less than N/2 samples, i.e. less than 50%, and of frames for which are delivered incomplete sets of synthesis parameters.

The frames for which complete sets of synthesis parameters are not delivered may form the subject of no spectral analysis. As a variant, an analysis may nevertheless be performed for these frames, so as to deliver incomplete sets of synthesis parameters including data representing an error of interpolation of at least one of the synthesis parameters and/or data representing a filter for interpolating at least one of the synthesis parameters.

In a first field of application of the method, the processing of the spectrum of the audio signal comprises an extraction of coding parameters with a view to the transmission and/or the storage of the coded audio signal. In a second field of application of the method, the processing of the spectrum of the audio signal comprises a denoising by spectral subtraction. Other fields of application may also be envisaged among audio processings.

A second aspect of the invention relates to a method of synthesizing an audio signal, in which successive spectral estimates corresponding respectively to frames of N samples of the audio signal which are weighted by an analysis window are obtained, the successive frames exhibiting mutual overlaps of L samples, each frame of the audio signal is evaluated by transforming the spectral estimates in the time domain, and the frames evaluated are combined to form the synthesized signal. According to this method, each evaluated frame is modified by applying thereto a processing corresponding to a division by said analysis window and to a

multiplication by a synthesis window, and the synthesized signal is formed as an overlap sum of the modified frames. The number L being smaller than N/2 and the samples of a frame having ranks i numbered from 0 to N-1, the synthesis window $f_S(i)$ satisfies $f_S(N-L+i) + f_S(i) = A$ for $0 \leq i < L$, and is equal to A for $L \leq i < N-L$, A being a positive constant.

In a variant of the synthesis method according to the invention, a set of successive overlapping frames of N samples of the audio signal which are weighted by an analysis window is evaluated, by transforming in the time domain spectral estimates corresponding respectively to said frames, and the evaluated frames are combined to form the synthesized signal. For a subset of the evaluated frames, the spectral estimates are obtained by processing synthesis parameters respectively associated with the frames of said subset while, for the frames which do not form part of the subset, the spectral estimates are obtained with an interpolation of a part at least of the synthesis parameters. The successive frames of said subset exhibit mutual time shifts of M samples, the number M being larger than N/2, while the successive frames of said set exhibit mutual time shifts of M/p samples, p being an integer larger than 1. Each evaluated frame is modified by applying thereto a processing corresponding to a division by said analysis window and to a multiplication by a synthesis window, and the synthesized signal is formed as an overlap sum of the modified frames. The samples of a frame having ranks i numbered from 0 to N-1, the synthesis window $f'_S(i)$ has a support limited to the ranks i ranging from $N/2 - M/p$ to $N/2 + M/p$ and satisfies $f'_S(i) + f'_S(i + M/p) = A$ for $N/2 - M/p \leq i < N/2$, A being a positive constant.

The invention also proposes audio processing devices comprising means for implementing the hereinabove methods of analysis and synthesis.

Other features and advantages of the present invention will become apparent in the description below of non-limiting exemplary embodiments, with reference to the appended drawings, in which:

- figure 1 is a schematic diagram of an audio coder according to the invention;
- figures 2 and 3 are charts illustrating the formation of the audio signal frames in the coder of figure 1;
- figures 4 and 5 are graphs showing an exemplary spectrum of the audio signal and illustrating the extraction of the upper and lower envelopes of this spectrum;
- figure 6 is a schematic diagram of an example of quantization means usable in the coder of figure 1;
- figure 7 is a schematic diagram of means usable to extract parameters relating to the phase of the non-harmonic component in a variant of the coder of figure 1;
- figure 8 is a schematic diagram of an audio decoder corresponding to the coder of figure 1;
- figure 9 is a flowchart of an exemplary procedure for smoothing spectral coefficients and for extracting minimum phases implemented in the decoder of figure 8;
- figure 10 is a schematic diagram of modules for analysis and for spectral mixing of harmonic and non-harmonic components of the audio signal;
- figures 11 to 13 are graphs showing examples of nonlinear functions usable in the analysis module of figure 10;

- figures 14 and 15 are charts illustrating a way of carrying out the temporal synthesis of the signal frames in the decoder of figure 8;
- figures 16 and 17 are graphs showing windowing functions usable in the synthesis of the frames according to figures 14 and 15;
- figures 18 and 19 are schematic diagrams of interpolation means usable in a variant embodiment of the coder and of the decoder;
- figure 20 is a schematic diagram of interpolation means usable in another variant embodiment of the coder; and
- figures 21 and 22 are charts illustrating another way of carrying out the temporal synthesis of the signal frames in the decoder of figure 8, with the aid of an interpolation of parameters.

The coder and decoder described hereinbelow are digital circuits which can, as is customary in the field of audio signal processing, be embodied by programming a digital signal processor (DSP) or an application specific integrated circuit (ASIC).

The audio coder represented in figure 1 processes an audio input signal x which, in the nonlimiting example considered hereinbelow, is a speech signal. The signal x is available in digital form, for example at a sampling frequency F_e of 8 kHz. It is, for example, delivered by an analog/digital converter processing the amplified output signal from a microphone. The input signal x can also be formed from another version, analog or digital, coded or uncoded, of the speech signal.

The coder comprises a module 1 which forms successive frames of audio signal for the various processing operations performed, and an output multiplexer 6 which delivers an output stream Φ containing, for each frame, sets of quantization parameters from which a decoder

will be capable of synthesizing a decoded version of the audio signal.

The structure of the frames is illustrated by figures 2 and 3. Each frame 2 is composed of a number N of consecutive samples of the audio signal x. The successive frames exhibit mutual time shifts corresponding to M samples, so that their overlap is L = N-M samples of the signal. In the example considered, where N = 256, M = 160 and L = 96, the duration of the frames 2 is N/F_e = 32 ms, and a frame is formed every M/F_e = 20 ms.

In a conventional manner, the module 1 multiplies the samples of each frame 2 by a windowing function f_A, preferably chosen for its good spectral properties. The samples x(i) of the frame being digitized from i = 0 to i = N-1, the analysis window $f_A(i)$ can thus be a Hamming window, expressed by:

$$f_A(i) = 0.54 + 0.46.\cos\left(2\pi\,\frac{i-(N-1)/2}{N}\right) \qquad (1)$$

or a Hanning window, expressed by:

$$f_A(i) = \frac{1}{2}\left(1+\cos\left(2\pi\,\frac{i-(N-1)/2}{N}\right)\right) \qquad (2)$$

or else a Kaiser window, expressed by:

$$f_A(i) = \frac{I_0\left(\alpha\sqrt{1-\left(\frac{i-(N-1)/2}{N}\right)^2}\right)}{I_0(\alpha)} \qquad (3)$$

where α is a coefficient equal, for example, to 6, and $I_0(.)$ designates the Bessel function of index 0.

The coder of figure 1 carries out an analysis of the audio signal in the spectral domain. It comprises a module 3 which calculates the fast Fourier transform (FFT) of each signal frame. The signal frame is shaped before being subjected to the FFT module 3: the module 1 appends N = 256 zero samples thereto so as to obtain the maximum resolution of the Fourier transform, and it

moreover performs a circular permutation of the $2N = 512$ samples so as to compensate for the phase effects resulting from the analysis window. This modification of the frame is illustrated by figure 3. The frame whose fast Fourier transform is calculated on $2N = 512$ points commences with the last $N/2 = 128$ weighted samples of the frame, followed by the $N = 256$ zero samples, and terminates with the first $N/2 = 128$ weighted samples of the frame.

The FFT module 3 obtains the spectrum of the signal for each frame, whose modulus and phase are respectively denoted $|X|$ and φ_X, or $|X(i)|$ and $\varphi_X(i)$ for the frequency indices $i = 0$ to $i = 2N-1$ (by virtue of the symmetry of the Fourier transform and of the frames, we may confine ourselves to the values for $0 \leq i < N$).

A fundamental-frequency detector 4 estimates for each signal frame a value of the fundamental frequency F_0. The detector 4 can apply any known procedure for analyzing the speech signal of the frame to estimate the fundamental frequency F_0, for example a procedure based on the autocorrelation function or the AMDF function, possibly preceded by a module for whitening by linear prediction. The estimate can also be made in the spectral domain or in the cepstral domain. Another possibility is to evaluate the time intervals between the consecutive breaks in the speech signal which are attributable to closures of the talker's glottis occurring over the duration of the frame. Well-known procedures which can be used to detect such microbreaks are described in the following articles: M. Basseville et al., "Sequential detection of abrupt changes in spectral characteristics of digital signals" (IEEE Trans. on Information Theory, 1983, Vol. IT-29, No. 5, pages 708-723); R. Andre-Obrecht, "A new statistical approach for the automatic segmentation of continuous speech signals" (IEEE Trans. on Acous., Speech and Sig. Proc., Vol. 36, No. 1, January 1988); and C. MURGIA et

al., "An algorithm for the estimation of glottal closure instants using the sequential detection of abrupt changes in speech signals" (Signal Processing VII, 1994, pages 1685-1688).

The estimated fundamental frequency F_0 forms the subject of a quantization, for example scalar, by a module 5, which provides the output multiplexer 6 with an index iF of quantization of the fundamental frequency for each frame of the signal.

The coder uses cepstral parametric modelings to represent an upper envelope and a lower envelope of the spectrum of the audio signal. The first step of the cepstral transformation consists in applying a spectral compression function to the modulus of the spectrum of the signal, which function may be a logarithmic or root function. The module 8 of the coder thus carries out, for each value X(i) of the spectrum of the signal ($0 \leq i < N$), the following transformation:

$$LX(i) = \mathrm{Log}(|X(i)|) \tag{4}$$

in the case of a logarithmic compression or

$$LX(i) = |X(i)|^{\gamma} \tag{5}$$

in the case of a root compression, γ being an exponent lying between 0 and 1.

The compressed spectrum LX of the audio signal is processed by a module 9 which extracts spectral amplitudes associated with the harmonics of the signal corresponding to the multiples of the estimated fundamental frequency F0. These amplitudes are then interpolated by a module 10 so as to obtain a compressed upper envelope denoted LX_sup.

It should be noted that the spectral compression could equivalently be performed after determining the amplitudes associated with the harmonics. It could also be performed after interpolation, and this would merely modify the form of the interpolation functions.

The module 9 for extracting the maxima takes account of any variation in the fundamental frequency over the analysis frame, errors which the detector 4 may make, as well as inaccuracies related to the discrete nature of the frequency sampling. To do this, the search for the amplitudes of the spectral peaks does not consist simply in taking the values LX(i) corresponding to the indices i such that $i.F_e/2N$ is the frequency closest to a harmonic of frequency $k.F_0 (k \geq 1)$. The spectral amplitude retained for a harmonic of order k is a local maximum of the modulus of the spectrum in the neighborhood of the frequency $k.F_0$ (this amplitude is obtained directly in compressed form when the spectral compression 8 is performed before the extraction of the maxima 9).

Figures 4 and 5 show an exemplary form of the compressed spectrum LX, where it may be seen that the maximum amplitudes of the harmonic peaks do not necessarily coincide with the amplitudes corresponding to the integer multiples of the estimated fundamental frequency F_0. Since the sides of the peaks are fairly steep, a small error in the positioning of the fundamental frequency F_0, amplified by the harmonic index k, may greatly distort the estimated upper envelope of the spectrum and cause poor modeling of the formant structure of the signal. For example, directly taking the spectral amplitude for the frequency $3.F_0$ in the case of figures 4 and 5 would produce a sizeable error in the extraction of the upper envelope in the neighborhood of the harmonic of order k = 3, although, in the example drawn, this relates to a zone of sizeable energy. By performing the interpolation on the basis of the actual maximum, this kind of error in estimating the upper envelope is avoided.

In the example represented in figure 4, the interpolation is performed between points whose

abscissa is the frequency corresponding to the maximum of the amplitude of a spectral peak, and whose ordinate is this maximum, before or after compression.

The interpolation performed to calculate the upper envelope LX_sup is a simple linear interpolation. Of course, some other form of interpolation could be used (for example polynomial or spline).

In the preferred variant represented in figure 5, the interpolation is performed between points whose abscissa is a frequency $k.F_0$ which is a multiple of the fundamental frequency (in fact the closest frequency in the discrete spectrum) and whose ordinate is the maximum amplitude, before or after compression, of the spectrum in the neighborhood of this multiple frequency.

By comparing figures 4 and 5, it may be seen that the mode of extraction according to figure 5, which repositions the peaks on the harmonic frequencies, leads to better accuracy with regard to the amplitude of the peaks which will be attributed by the decoder to the frequencies which are multiples of the fundamental frequency. A slight frequency displacement may occur in the position of these peaks, this not being very significant perceptually and anyway not being avoided either in the case of figure 4. In the case of figure 4, the anchoring points for the interpolation are one and the same as the vertices of the harmonic peaks. In the case of figure 5, these anchoring points must lie precisely at the frequencies which are multiples of the fundamental frequency, their amplitudes corresponding to those of the peaks.

The search interval for the amplitude maximum associated with a harmonic of rank k is centered on the index i of the frequency of the FFT closest to $k.F_0$,

i.e. $i = \left\lfloor 2Nk\frac{F_0}{F_e} + \frac{1}{2} \right\rfloor$, where $\lfloor a \rfloor$ designates the integer equal to or immediately less than the number a. The width of this search interval depends on the sampling frequency F_e, on the size 2N of the FFT and on the possible range of variation of the fundamental frequency. This width is typically of the order of some ten frequencies with the exemplary values considered earlier. It may be rendered adjustable as a function of the value F_0 of the fundamental frequency and of the number k of the harmonic.

In order to improve the resolution in the low frequencies and hence to more faithfully represent the amplitudes of the harmonics in this zone, a nonlinear distortion of the frequency scale is carried out on the compressed upper envelope by a module 12 before the module 13 performs the inverse fast Fourier transform (IFFT) providing the cepstral coefficients cx_sup.

The nonlinear distortion allows more efficient minimization of the modeling error. It is, for example, performed on a frequency scale of Mel or Bark type. This distortion may possibly depend on the estimated fundamental frequency F_0. Figure 1 illustrates the case of the Mel scale. The relation between the frequencies F of the linear spectrum, expressed in hertz, and the frequencies F′ of the Mel scale is as follows:

$$F' = \frac{1000}{\log_{10}(2)} \times \log_{10}\left(1 + \frac{F}{1000}\right) \qquad (6)$$

In order to limit the transmission bit rate, a truncation of the cepstral coefficients cx_sup is performed. The IFFT module 13 need only calculate a cepstral vector of NCS cepstral coefficients of orders 0 to NCS-1. By way of example, NCS may be equal to 16.

Post-filtering in the cepstral domain, referred to as post-liftering, is applied by a module 15 to the compressed upper envelope LX_sup. This post-liftering corresponds to a manipulation of the cepstral coefficients cx_sup delivered by the IFFT module 13, which corresponds approximately to a post-filtering of the harmonic part of the signal by a transfer function having the conventional form:

$$H(z) = \left(1 - \mu z^{-1}\right)\frac{A(z / \gamma_1)}{A(z / \gamma_2)} \qquad (7)$$

where A(z) is the transfer function of a filter for linear prediction of the audio signal, γ_1 and γ_2 are coefficients lying between 0 and 1, and μ is a pre-emphasizing coefficient, possibly zero. The relation between the post-liftered coefficient of order i, denoted $c_p(i)$, and the corresponding cepstral coefficient c(i) = cx_sup(i) delivered by the module 13 is then:

$$\begin{aligned} c_p(0) &= c(0) \\ c_p(i) &= \left(1 + \gamma_2^i - \gamma_1^i\right)c(i) - \frac{\mu^i}{i} \quad \text{for } i > 0 \end{aligned} \qquad (8)$$

The optional pre-emphasizing coefficient μ may be controlled by setting as constraint the preserving of the value of the cepstral coefficient cx_sup(1) relating to the slope. Specifically, the value of c(1) = cx_sup(1) of white noise filtered by the pre-emphasizing filter corresponds to the pre-emphasizing coefficient. The latter may thus be chosen as follows: $\mu = (\gamma_2-\gamma_1).c(1)$.

After the post-lifter 15, a normalizing module 16 again modifies the cepstral coefficients by imposing the constraint of exact modeling of a point of the initial spectrum, which is preferably the point of greatest energy from among the spectral maxima extracted by the module 9. In practice, this normalization modifies only the value of the coefficient $c_p(0)$.

The normalizing module 16 operates as follows: it recalculates a value of the synthesized spectrum at the frequency of the maximum indicated by the module 9, by Fourier transform of the truncated and post-liftered cepstral coefficients, taking into account the nonlinear distortion of the frequency axis; it determines a normalizing gain g_N through the logarithmic difference between the value of the maximum as delivered by the module 9 and this value recalculated; and it adds the gain g_N to the post-liftered cepstral coefficient $c_p(0)$. This normalization may be viewed as being part of the post-liftering.

The post-liftered and normalized cepstral coefficients form the subject of a quantization by a module 18 which transmits corresponding quantization indices icxs to the output multiplexer 6 of the coder.

The module 18 can operate by vector quantization on the basis of cepstral vectors formed of post-liftered and normalized coefficients, here denoted cx[n] for the signal frame of rank n. By way of example, the cepstral vector cx[n] of NCS = 16 cepstral coefficients cx[n,0], cx[n,1], ..., cx[n,NCS-1] is distributed as four cepstral subvectors each containing four coefficients of consecutive orders. The cepstral vector cx[n] can be processed by the means represented in figure 6, forming part of the quantization module 18. These means implement, for each component cx[n,i], a predictor of the form:

$$cx_p[n,i] = (1-\alpha(i)).rcx[n,i] + \alpha(i).rcx[n-1,i] \quad (9)$$

where rcx[n] designates a residual prediction vector for the frame of rank n whose components are respectively denoted rcx[n,0], rcx[n,1], ..., rcx[n,NCS-1], and $\alpha(i)$ designates a prediction coefficient chosen so as to be representative of an assumed inter-frame correlation. After quantization of the residuals, this residual vector is defined by:

$$rcx[n, i] = \frac{cx[n, i] - \alpha(i) .rcx\_q[n-1, i]}{2 - \alpha(i)} \qquad (10)$$

where rcx_q[n-1] designates the quantized residual vector for the frame of rank n-1, whose components are respectively denoted rcx_q[n,0], rcx_q[n,1], ..., rcx_q[n,NCS-1].

The numerator of relation (10) is obtained by a subtractor 20, whose output vector components are divided by the quantities 2-α(i) at 21. For quantization purposes, the residual vector rcx[n] is subdivided into four subvectors, corresponding to the subdivision into four cepstral subvectors. On the basis of a dictionary obtained by prior learning, the unit 22 undertakes the vector quantization of each subvector of the residual vector rcx[n]. This quantization can consist, for each subvector srcx[n], in selecting from the dictionary the quantized subvector srcx_q[n] which minimizes the quadratic error $\|srcx[n] - srcx\_q[n]\|^2$. The set icxs of quantization indices icx, corresponding to the addresses in the dictionary or dictionaries of the quantized residual subvectors srcx_q[n], is provided to the output multiplexer 6.

The unit 22 also delivers the values of the quantized residual subvectors, which form the vector rcx_q[n]. The latter is delayed by one frame at 23, and its components are multiplied by the coefficients α(i) at 24 so as to provide the vector to the negative input of the subtractor 20. The latter vector is, on the other hand, provided to an adder 25, the other input of which receives a vector formed by the components of the quantized residual rcx_q[n], respectively multiplied by the quantities 1-α(i) at 26. The adder 25 thus delivers the quantized cepstral vector cx_q[n] which will be recovered by the decoder.

The prediction coefficient $\alpha(i)$ can be optimized separately for each of the cepstral coefficients. The quantization dictionaries may also be optimized separately for each four cepstral subvectors. Moreover, it is possible, in a manner known per se, to normalize the cepstral vectors before applying the prediction/quantization scheme, on the basis of the variance of the cepstra.

It should be noted that the above scheme for quantizing the cepstral coefficients cannot be applied other than in respect of certain of the frames only. For example, provision may be made for a second mode of quantization as well as a process for selecting that one of the two modes which minimizes a least squares criterion with the cepstral coefficients to be quantized, and a bit indicating which of the two modes has been selected may be transmitted with the frame quantization indices.

The quantized cepstral coefficients cx_sup_q = cx_q[n] provided by the adder 25 are addressed to a module 28 which recalculates the spectral amplitudes associated with one or more of the harmonics of the fundamental frequency F_0 (figure 1). These spectral amplitudes are, for example, calculated in compressed form, by applying the Fourier transform to the quantized cepstral coefficients while taking account of the nonlinear distortion of the frequency scale used in the cepstral transformation. The amplitudes thus recalculated are provided to an adaptation module 29 which compares them with amplitudes of maxima determined by the extraction module 9.

The adaptation module 29 controls the post-lifter 15 in such a way as to minimize a discrepancy in modulus between the spectrum of the audio signal and the corresponding modulus values calculated at 28. This discrepancy in modulus can be expressed by a sum of absolute values of differences of amplitudes,

compressed or otherwise, corresponding to one or more of the harmonic frequencies. This sum can be weighted as a function of the spectral amplitudes associated with these frequencies.

Optimally, the discrepancy in modulus taken into account in the adaptation of the post-liftering would take account of all the harmonics of the spectrum. However, in order to reduce the complexity of the optimization, the module 28 can resynthesize the spectral amplitudes for just one or more frequencies which are multiples of the fundamental frequency F_0 and which are selected on the basis of the magnitude of the modulus of the spectrum in absolute value. The adaptation module 29 can, for example, consider the three most intense spectral peaks in the calculation of the discrepancy in modulus to be minimized.

In another embodiment, the adaptation module 29 estimates a curve of spectral masking of the audio signal by means of a psycho-acoustic model, and the frequencies taken into account in the calculation of the discrepancy in modulus to be minimized are selected on the basis of the magnitude of the modulus of the spectrum in relation to the masking curve (it is, for example, possible to take the three frequencies for which the modulus of the spectrum most exceeds the masking curve). Various conventional methods can be used to calculate the masking curve from the audio signal. It is, for example, possible to use that developed by J.D. Johnston ("Transform Coding of Audio Signals Using Perceptual Noise Criteria", IEEE Journal on Selected Area in Communications, Vol. 6, No. 2, February 1988).

To carry out the adaptation of the post-liftering, the module 29 can use a filter identification model. A simpler method consists in predefining a collection of sets of post-liftering parameters, that is to say a

collection of pairs γ_1, γ_2 in the case of post-liftering according to relations (8), in performing the operations incumbent on the modules 15, 16, 18 and 28 for each of these sets of parameters, and in retaining that of the sets of parameters which leads to the minimum discrepancy in modulus between the spectrum of the signal and the recalculated values. The quantization indices provided by the module 18 are then those which relate to the best set of parameters.

By a process similar to that for extracting the coefficients cx_sup representing the compressed upper envelope LX_sup of the spectrum of the signal, the coder determines the coefficients cx_inf representing a compressed lower envelope LX_inf. A module 30 extracts from the compressed spectrum LX, spectral amplitudes associated with frequencies situated in zones of the spectrum which are intermediate with respect to the frequencies which are multiples of the estimated fundamental frequency F_0.

In the example illustrated by figures 4 and 5, each amplitude associated with a frequency situated in a zone intermediate between two successive harmonics $k.F_0$ and $(k+1).F_0$ corresponds simply to the modulus of the spectrum for the frequency $(k+1/2).F_0$ situated in the middle of the interval separating the two harmonics. In another embodiment, this amplitude could be an average of the modulus of the spectrum over a small span surrounding this frequency $(k+1/2).F_0$.

A module 31 carries out an interpolation, for example linear, of the spectral amplitudes associated with the frequencies situated in the intermediate zones so as to obtain the compressed lower envelope LX_inf.

The cepstral transformation applied to this compressed lower envelope LX_inf is performed according to a frequency scale resulting from a nonlinear distortion

applied by a module 32. The IFFT module 33 calculates a cepstral vector of NCI cepstral coefficients cx_inf of orders 0 to NCI-1 representing the lower envelope. NCI is a number which may be substantially smaller than NCS, for example NCI = 4.

The nonlinear transformation of the frequency scale for the cepstral transformation of the lower envelope can be carried out to a scale which is finer at the high frequencies than at the low frequencies, thereby advantageously allowing good modeling of the unvoiced components of the signal at the high frequencies. However, to ensure homogeneity of representation between the upper envelope and the lower envelope, the same scale will preferably be adopted in the module 32 as in the module 12 (Mel in the example considered).

The cepstral coefficients cx_inf representing the compressed lower envelope are quantized by a module 34, which may operate in the same manner as the module 18 for quantizing the cepstral coefficients representing the compressed upper envelope. In the case considered, where we restricted ourselves to NCI = 4 cepstral coefficients for the lower envelope, the vector thus formed is subjected to a prediction residual vector quantization performed by means identical to those represented in figure 6 but without subdivision into subvectors. The quantization index icx = icxi determined by the vector quantizer 22 for each frame relating to the coefficients cx_inf is provided to the output multiplexer 6 of the coder.

The coder represented in figure 1 does not comprise any particular device for coding the phases of the spectrum at the harmonics of the audio signal.

On the other hand, it comprises means 36-40 for coding time information related to the phase of the

nonharmonic component represented by the lower envelope.

A spectral decompression module 36 and an IFFT module 37 form a temporal estimate of the frame of the non-harmonic component. The module 36 applies a decompression function which is the reciprocal of the compression function applied by the module 8 (that is to say an exponential or a $1/\gamma$ power function) to the compressed lower envelope LX_inf produced by the interpolation module 31. This provides the modulus of the estimated frame of the nonharmonic component, whose phase is taken equal to that φ_X of the spectrum of the signal X over the frame. The inverse Fourier transform performed by the module 37 provides the estimated frame of the nonharmonic component.

The module 38 subdivides this estimated frame of the nonharmonic component into several time segments. The frame delivered by the module 37 being made up of $2N = 512$ weighted samples, as illustrated by figure 3, the module 38 considers only the first $N/2 = 128$ samples and the last $N/2 = 128$ samples, and subdivides them, for example, into eight segments of 32 consecutive samples each representing 4 ms of signal.

For each segment, the module 38 calculates the energy equal to the sum of the squares of the samples, and forms a vector E1 formed of eight positive real components equal to the eight calculated energies. The largest of these eight energies, denoted EM, is also determined so as to be provided, with the vector E1, to a normalizing module 39. The latter divides each component of the vector E1 by EM, so that the normalized vector Emix is formed of eight components lying between 0 and 1. It is this normalized vector Emix, or weighting vector, which is subjected to the quantization by the module 40. The latter can carry out a vector quantization with a dictionary determined

during prior learning. The quantization index iEm is provided by the module 40 to the output multiplexer 6 of the coder.

Figure 7 shows a variant embodiment of the means employed by the coder of figure 1 to determine the energy weighting vector Emix for the frame of the non-harmonic component. The spectral decompression and IFFT modules 36, 37 operate like those which bear the same references in figure 1. A selection module 42 is added so as to determine the value of the modulus of the spectrum subjected to the inverse Fourier transform 37. On the basis of the estimated fundamental frequency F_0, the module 42 identifies harmonic regions and non-harmonic regions of the spectrum of the audio signal. For example, a frequency will be regarded as belonging to a harmonic region if it is located in a frequency interval centered on a harmonic $k.F_0$ and of width corresponding to a synthesized spectral line width, and to a nonharmonic region otherwise. In the nonharmonic regions, the complex signal subjected to the IFFT 37 is equal to the value of the spectrum, that is to say its modulus and its phase correspond to the values $|X|$ and φ_X provided by the FFT module 3. In the harmonic regions, this complex signal has the same phase φ_X as the spectrum and a modulus given by the lower envelope after spectral decompression 36. Proceeding thus according to figure 7 achieves more accurate modeling of the nonharmonic regions.

The decoder represented in figure 8 comprises an input demultiplexer 45 which extracts from the binary stream Φ, emanating from a coder according to figure 1, the quantization indices iF, icxs, icxi, iEm for the fundamental frequency F_0, the cepstral coefficients representing the compressed upper envelope, the coefficients representing the compressed lower envelope, and the weighting vector Emix, and distributes them respectively to modules 46, 47, 48 and

49. These modules 46-49 comprise quantization dictionaries similar to those of the modules 5, 18, 34 and 40 of figure 1, so as to restore the values of the quantized parameters. The modules 47 and 48 have dictionaries so as to form the quantized prediction residuals rcx_q[n], and they deduce therefrom the quantized cepstral vectors cx_q[n] with elements identical to the elements 23-26 of figure 6. These quantized cepstral vectors cx_q[n] provide the cepstral coefficients cx_sup_q and cx_inf_q processed by the decoder.

A module 51 calculates the fast Fourier transform of the cepstral coefficients cx_sup for each signal frame. The frequency scale of the compressed spectrum resulting therefrom is modified nonlinearly by a module 52 applying the nonlinear transformation reciprocal to that of the module 12 of figure 1, and which provides the estimate LX_sup of the compressed upper envelope. A spectral decompression of LX_sup, carried out by a module 53, provides the upper envelope X_sup comprising the estimated values of the modulus of the spectrum at the frequencies which are multiples of the fundamental frequency F_0. The module 54 synthesizes the spectral estimate X_v of the harmonic component of the audio signal, through a sum of spectral lines centered on the frequencies which are multiples of the fundamental frequency F_0 and whose amplitudes (in modulus) are those given by the upper envelope X_sup.

Although the digital input stream Φ does not comprise any specific information regarding the phase of the spectrum of the signal at the harmonics of the fundamental frequency, the decoder of figure 8 is capable of extracting information regarding this phase from the cepstral coefficients cx_sup_q representing the compressed upper envelope. This phase information is used to assign a phase $\varphi(k)$ to each of the spectral

lines determined by the module 54 in the estimate of the harmonic component of the signal.

As a first approximation, the speech signal may be regarded as being of minimum phase. Moreover, it is known that the minimum phase information may be deduced easily from cepstral modeling. This minimum phase information is therefore calculated for each harmonic frequency. The minimum phase assumption signifies that the energy of the synthesized signal is localized at the start of each period of the fundamental frequency F_0.

In order to be closer to a real speech signal, slight dispersion is introduced by means of a specific post-liftering of the cepstra during synthesis of the phase. With this post-liftering, performed by the module 55 of figure 8, it is possible to emphasize the formant resonances of the envelope and hence to control the dispersion of the phases. This post-liftering is, for example, of the form (8).

To limit the phase breaks, it is preferable to smooth the post-liftered cepstral coefficients, this being performed by the module 56. The module 57 deduces from the post-liftered and smoothed cepstral coefficients the minimum phase assigned to each spectral line representing a harmonic peak of the spectrum.

The operations performed by the modules 56, 57 for smoothing and extracting the minimum phase are illustrated by the flowchart of figure 9. The module 56 examines the variations in the cepstral coefficients so as to apply lesser smoothing in the presence of abrupt variations than in the presence of slow variations. To do this, it performs the smoothing of the cepstral coefficients by means of a forget factor λ_c chosen as a function of a comparison between a threshold d_{th} and a distance d between two successive sets of post-liftered

cepstral coefficients. The threshold d_{th} is itself adapted as a function of the variations of the cepstral coefficients.

The first step 60 consists in calculating the distance d between the two successive vectors relating to frames n-1 and n. These vectors, here denoted cxp[n-1] and cxp[n], correspond for each frame to the collection of NCS post-liftered cepstral coefficients representing the compressed upper envelope. The distance used may in particular be the Euclidean distance between the two vectors or else a quadratic distance.

Two smoothings are firstly performed, respectively by means of forget factors λ_{min} and λ_{max}, so as to determine a minimum distance d_{min} and a maximum distance d_{max}. The threshold d_{th} is then determined in step 70 as being situated between the minimum and maximum distances d_{min}, d_{max}: $d_{th} = \beta . d_{max} + (1-\beta) . d_{min}$, the coefficient β being, for example, equal to 0.5.

In the example represented, the forget factors λ_{min} and λ_{max} are themselves selected from among two distinct values, respectively λ_{min1}, λ_{min2} and λ_{max1}, λ_{max2} lying between 0 and 1, the indices λ_{min1}, λ_{max1} each being substantially nearer to 0 than the indices λ_{min2}, λ_{max2}. If $d > d_{min}$ (test 61), the forget factor λ_{min} is equal to λ_{min1} (step 62); otherwise, it is taken equal to λ_{min2} (step 63). In step 64, the minimum distance d_{min} is taken equal to $\lambda_{min} . d_{min} + (1-\lambda_{min}) . d$. If $d > d_{max}$ (test 65), the forget factor λ_{max} is equal to λ_{max1} (step 66); otherwise, it is taken equal to λ_{max2} (step 67). In step 68, the minimum distance d_{max} is taken equal to $\lambda_{max} . d_{max} + (1-\lambda_{max}) . d$.

If the distance d between the two consecutive cepstral vectors is greater than the threshold d_{th} (test 71), then a value λ_{c1} relatively close to 0 is adopted for the forget factor λ_c (step 72). In this case, the

corresponding signal is regarded as being of nonstationary type, so that there is no need to keep a large memory of the earlier cepstral coefficients. If $d \leq d_{th}$, a value λ_{c2} which is not as close to 0 is adopted in step 73 for the forget factor λ_c, so as to further smooth the cepstral coefficients. The smoothing is performed in step 74, where the vector cxl[n] of smoothed coefficients for the current frame n is determined by:

$$cxl[n] = \lambda_c . cxl[n-1] + (1-\lambda_c) . cxp[n] \qquad (11)$$

The module 57 then calculates the minimum phases $\varphi(k)$ associated with the harmonics $k.F_0$. In a known manner, the minimum phase for a harmonic of order k is given by:

$$\varphi(k) = -2 . \sum_{m=1}^{NCS-1} cxl[n, m] . \sin(2\pi mk\, F_0 / F_e) \qquad (12)$$

where cxl[n,m] designates the smoothed cepstral coefficient of order m for frame n.

In step 75, the harmonic index k is initialized to 1. To initialize the calculation of the minimum phase assigned to harmonic k, the phase $\varphi(k)$ and the cepstral index m are initialized to 0 and 1 respectively in step 76. In step 77, the module 57 adds the quantity $-2.cxl[n,m].\sin(2\pi mk.F_0/F_e)$ to the phase $\varphi(k)$. The cepstral index m is incremented in step 78 and compared with NCS in step 79. Steps 77 and 78 are repeated so long as m < NCS. When m = NCS, the calculation of the minimum phase is terminated for harmonic k, and the index k is incremented in step 80. The calculation of minimum phases 76-79 is rerun for the next harmonic so long as $k.F_0 < F_e/2$ (test 81).

In the exemplary embodiment according to figure 8, the module 54 takes account of a constant phase over the width of each spectral line, equal to the minimum phase $\varphi(k)$ provided for the corresponding harmonic k by the module 57.

The estimate X_v of the harmonic component is synthesized by summation of spectral lines positioned at the harmonic frequencies of the fundamental frequency F_0. During this synthesis, it is possible to position the spectral lines on the frequency axis with a higher resolution than the resolution of the Fourier transform. To do this, a reference spectral line is precalculated once and for all according to the higher resolution. This calculation can consist of a Fourier transform of the analysis window F_A with a transform size of 16 384 points, achieving a resolution of 0.5 Hz per point. The synthesis of each harmonic line is then performed by the module 54 by positioning on the frequency axis the reference line with high resolution, and by undersampling this reference spectral line so as to reduce to the resolution of 16.625 Hz of the Fourier transform on 512 points. This enables the spectral line to be positioned accurately.

For the determination of the lower envelope, the FFT module 85 of the decoder of figure 8 receives the NCI quantized cepstral coefficients cx_inf_q of orders 0 to NCI - 1, and it advantageously supplements them with the NCS - NCI cepstral coefficients cx_sup_q of order NCI to NCS - 1 representing the upper envelope. Specifically, it may be estimated that, as a first approximation, the fast variations of the compressed lower envelope are well reproduced by those of the compressed upper envelope. In another embodiment, the FFT module 85 could consider only the NCI cepstral parameters cx_inf_q.

The module 86 converts the frequency scale in a manner reciprocal to the conversion carried out by the module 32 of the coder, so as to restore the estimate LX_inf of the compressed lower envelope, subjected to the spectral decompression module 87. At the output of the module 87, the decoder is furnished with a lower

envelope X_inf comprising the values of the modulus of the spectrum in the valleys situated between the harmonic peaks.

This envelope X_inf will modulate the spectrum of a noise frame whose phase is processed as a function of the quantized weighting vector Emix extracted by the module 49. A generator 88 delivers a normalized noise frame whose 4-ms segments are weighted in a module 89 in accordance with the normalized components of the vector Emix provided by the module 49 for the current frame. This noise is white noise high-pass filtered so as to take account of the low level which in principle the unvoiced component has at the low frequencies. On the basis of the energy-weighted noise, the module 90 forms frames of 2N = 512 samples by applying the analysis window f_A, the insertion of 256 zero samples and the circular permutation for phase compensation in accordance with what was explained with reference to figure 3. The Fourier transform of the resulting frame is calculated by the FFT module 91.

The spectral estimate X_{uv} of the nonharmonic component is determined by the spectral synthesis module 92 which performs a frequency-by-frequency weighting. This weighting consists in multiplying each complex spectral value provided by the FFT module 91 by the value of the lower envelope X_inf obtained for the same frequency by the spectral decompression module 87.

The spectral estimates X_v, X_{uv} of the harmonic (voiced in the case of a speech signal) and nonharmonic (or unvoiced) components are combined by a mixing module 95 controlled by a module 96 for analyzing the degree of harmonicity (or of voicing) of the signal.

The organization of these modules 95, 96 is illustrated by figure 10. The analysis module 96 comprises a unit 97 for estimating a frequency-dependent degree of

voicing W from which are calculated four frequency-dependent gains, namely two gains g_v, g_{uv} controlling the relative magnitude of the harmonic and nonharmonic components in the synthesized signal, and two gains g_{v_φ}, g_{uv_φ} used to add noise to the phase of the harmonic component.

The degree of voicing W(i) is a continuously varying value lying between 0 and 1 determined for each frequency index i ($0 \le i < N$) as a function of the upper envelope X_sup(i) and of the lower envelope X_inf(i) which are obtained for this frequency i by the decompression modules 53, 87. The degree of voicing W(i) is estimated by the unit 97 for each frequency index i corresponding to a harmonic of the fundamental frequency F_0, namely $i = \left\lfloor 2Nk\frac{F_0}{F_e} + \frac{1}{2} \right\rfloor$ for k = 1, 2, ..., by an increasing function of the ratio of the upper envelope X_sup to the lower envelope X_inf at this frequency, for example according to the formula:

$$W(i) = \min\left\{1, \frac{10.\log_{10}[X\_\sup(i)/X\_\inf(i)]}{Vth(F_0)}\right\} \qquad (13)$$

The threshold $Vth(F_0)$ corresponds to the average dynamic swing calculated over a purely voiced synthetic spectrum at the fundamental frequency. It is advantageously chosen to be dependent on the fundamental frequency F_0.

The degree of voicing W(i) for a frequency other than the harmonic frequencies is obtained simply as being equal to that estimated for the closest harmonic.

The gain $g_v(i)$, which depends on the frequency, is obtained by applying a nonlinear function to the degree of voicing W(i) (block 98). This nonlinear function has, for example, the form represented in figure 11:

$$g_v(i) = 0 \text{ if } 0 \le W(i) \le W1$$

$$g_v(i) = \frac{W(i) - W1}{W2 - W1} \text{ if } W1 < W(i) < W2 \qquad (14)$$

$$g_v(i) = 1 \text{ if } W2 \le W(i) \le 1$$

the thresholds W1, W2 being such that $0 < W1 < W2 < 1$. The gain g_{uv} can be calculated in a similar manner to the gain g_v (the sum of the two gains g_v, g_{uv} being constant, for example equal to 1), or deduced simply from the latter through the relation $g_{uv}(i) = 1 - g_v(i)$, as shown diagrammatically by the subtractor 99 in figure 10.

It is beneficial to be able to add noise to the phase of the harmonic component of the signal at a given frequency if the analysis of the degree of voicing shows that the signal is actually of nonharmonic type at this frequency. To do this, the phase φ_v' of the mixed harmonic component is the result of a linear combination of the phases φ_v, φ_{uv} of the harmonic and nonharmonic components X_v, X_{uv} synthesized by the modules 54, 92.

The gains g_{v_φ}, g_{uv_φ} respectively applied to these phases are calculated from the degree of voicing W and also weighted as a function of the frequency index i, given that the adding of noise to the phase is actually useful only beyond a certain frequency.

A first gain g_{v1_φ} is calculated by applying a nonlinear function to the degree of voicing W(i), as shown diagrammatically by the block 100 in figure 10. This nonlinear function can have the form represented in figure 12:

$$g_{v1\_\varphi}(i) = G1 \text{ if } 0 \le W(i) \le W3$$

$$g_{v1\_\varphi}(i) = G1 + (1 - G1)\frac{W(i) - W3}{W4 - W3} \text{ if } W3 < W(i) < W4 \qquad (15)$$

$$g_{v1\_\varphi}(i) = 1 \text{ if } W4 \le W(i) \le 1$$

the thresholds W3 and W4 being such that $0 < W3 < W4 < 1$, and the minimum gain G1 lying between 0 and 1.

A multiplier 101 multiplies for each frequency of index i the gain g_{v1_φ} by another gain g_{v2_φ} dependent only on the frequency index i, so as to form the gain $g_{v\_\varphi}(i)$. The gain $g_{v2\_\varphi}(i)$ depends nonlinearly on the frequency index i, for example as indicated in figure 13:

$$g_{v2\_\varphi}(i) = 1 \quad \text{if } 0 \le i \le i1$$

$$g_{v2\_\varphi}(i) = 1 - (1 - G2)\frac{i - i1}{i2 - i1} \quad \text{if } i1 < i < i2 \qquad (16)$$

$$g_{v2\_\varphi}(i) = G2 \quad \text{if } i2 \le i \le 1$$

the indices i1 and i2 being such that $0 < i1 < i2 \le N$, and the minimum gain G2 lying between 0 and 1. The gain $g_{uv\_\varphi}(i)$ can be calculated simply as being equal to $1 - g_{v\_\varphi}(i) = 1 - g_{v1\_\varphi}(i).g_{v2\_\varphi}(i)$ (subtractor 102 of figure 10).

The complex spectrum Y of the synthesized signal is produced by the mixing module 95, which carries out the following mixing relation, for $0 \le i < N$:

$$Y(i) = g_v(i).|X_v(i)|.\exp[j\varphi'_v(i)]+g_{uv}(i).X_{uv}(i) \qquad (17)$$

with $$\varphi'_v(i) = g_{v\_\varphi}(i).\varphi_v(i) + g_{uv\_\varphi}(i).\varphi_{uv}(i) \qquad (18)$$

where $\varphi_v(i)$ designates the argument of the complex number $X_v(i)$ provided by the module 54 for the frequency of index i (block 104 of figure 10), and $\varphi_{uv}(i)$ designates the argument of the complex number $X_{uv}(i)$ provided by the module 92 (block 105 of figure 10). This combination is carried out by the multipliers 106-110 and the adders 111-112 represented in figure 10.

The mixed spectrum Y(i) for $0 \le i < 2N$ (with Y(2N-1-i) = Y(i)) is then transformed into the time domain by the IFFT module 115 (figure 8). Only the first N/2 = 128 and the last N/2 = 128 samples of the frame of 2N = 512 samples produced by the module 115 are retained, and the circular permutation inverse to that illustrated by figure 3 is applied to obtain the synthesized frame of N = 256 samples weighted by the analysis window f_A.

The frames obtained successively in this manner are finally processed by the temporal synthesis module 116 which forms the decoded audio signal $\hat{x}$.

The temporal synthesis module 116 performs an overlap sum of frames modified with respect to those evaluated successively at the output of the module 115. The modification may be viewed in two steps illustrated by figures 14 and 15 respectively.

The first step (figure 14) consists in multiplying each frame 2′ delivered by the IFFT module 115 by a window $1/f_A$ inverse to the analysis window f_A employed by the module 1 of the coder. The samples of the frame 2″ resulting therefrom are therefore uniformly weighted.

The second step (figure 15) consists in multiplying the samples of this frame 2″ by a synthesis window f_S satisfying the following properties:

$$f_S(N-L+i) + f_S(i) = A \quad \text{for } 0 \le i < L \qquad (19)$$

$$f_S(i) = A \quad \text{for } L \le i < N-L \qquad (20)$$

where A designates an arbitrary positive constant, for example A = 1. The synthesis window $f_S(i)$ increases progressively from 0 to A for i going from 0 to L. It is, for example, a raised half-sinusoid:

$$f_S(i) = \frac{A}{2}.(1 - \cos[(i + 1/2)\pi / L]) \quad \text{for } 0 \le i < L \quad (21)$$

After having reweighted each frame 2″ by the synthesis window f_S, the module 116 positions the successive frames with their time shifts of M = 160 samples and their time overlaps of L = 96 samples, then it sums the frames thus positioned over time. Owing to the properties (19) and (20) of the synthesis window f_S, each sample of the decoded audio signal $\hat{x}$ thus obtained is assigned a uniform global weight, equal to A. This global weight originates from the contribution of a single frame if the sample has in this frame a rank i such that $L \le i < N - L$, and comprises the summed

contributions of two successive frames if $0 \le i < L$ where $N - L \le i < N$.

It is thus possible to perform the temporal synthesis in a simple manner even if, as in the case considered, the overlap L between two successive frames is smaller than half the size N of these frames.

The two steps set forth above for modifying the signal frames may be merged into a single step. It is sufficient to precalculate a compound window $f_C(i) = f_S(i)/f_A(i)$ and simply to multiply the frames 2' of N = 256 samples delivered by the module 115 by the compound window f_C before performing the overlap summation.

Figure 16 shows the shape of the compound window f_C in the case where the analysis window f_A is a Hamming window and the synthesis window f_S has the form given by relations (19) to (21).

Other forms of the synthesis window f_S satisfying relations (19) and (20) may be employed. In the variant of figure 17, it is a piecewise affine function defined by:

$$f_S(i) = A.i/L \text{ for } 0 \le i < L \qquad (22)$$

In order to improve the quality of coding of the audio signal, the coder of figure 1 can increase the rate of formation and of analysis of the frames, so as to transmit more quantization parameters to the decoder. In the frame structure represented in figure 2, a frame of N = 256 samples (32 ms) is formed every 20 ms. These frames of 256 samples could be formed at a higher rate, for example 10 ms, two successive frames then having a shift of M/2 = 80 samples and an overlap of 176 samples.

Under these conditions, it is possible to transmit the complete sets of quantization parameters iF, icxs, icxi, iEm for just one subcollection of frames, and to transmit, for the other frames, parameters making it possible to perform a suitable interpolation at the level of the decoder. In the example envisaged hereinabove, the subcollection for which complete parameter sets are transmitted may consist of the frames of integer rank n, whose periodicity is M/F_e = 20 ms, and the frames for which an interpolation is performed may be those of half-integer rank n + 1/2 which are shifted by 10 ms with respect to the frames of the subcollection.

In the embodiment illustrated by figure 18, the notation cx_q[n-1] and cx_q[n] designates quantized cepstral vectors determined, for two successive frames of integer rank, by the quantization module 18 and/or by the quantization module 34. These vectors comprise, for example, four consecutive cepstral coefficients each. They could also comprise more cepstral coefficients.

A module 120 performs an interpolation of these two cepstral vectors cx_q[n-1] and cx_q[n] so as to estimate an intermediate value cx_i[n-1/2]. The interpolation performed by the module 120 can be a simple arithmetic average of the vectors cx_q[n-1] and cx_q[n]. As a variant, the module 120 could apply a more sophisticated interpolation formula, for example polynomial, based also on the cepstral vectors obtained for frames earlier than frame n-1. Moreover, if more than one interpolated frame is interposed between two consecutive frames of integer rank, the interpolation takes account of the relative position of each interpolated frame.

With the aid of the means described above, the coder also calculates the cepstral coefficients cx[n-1/2]

relating to the frame of half-integer rank. In the case of the upper envelope, these cepstral coefficients are those provided by the IFFT module 13 after post-liftering 15 (for example with the same post-liftering coefficients as for the previous frame n-1) and normalization 16. In the case of the lower envelope, the cepstral coefficients cx[n-1/2] are those delivered by the IFFT module 33.

A subtractor 121 forms the difference ecx[n-1/2] between the cepstral coefficients cx[n-1/2] calculated for the frame of half-integer rank and the coefficients cx_i[n-1/2] estimated by interpolation. This difference is provided to a quantization module 122 which addresses quantization indices icx[n-1/2] to the output multiplexer 6 of the coder. The module 122 operates, for example, by vector quantization of the interpolation errors ecx[n-1/2] determined successively for the frames of half-integer rank.

This quantization of the interpolation error can be performed by the coder for each of the NCS + NCI cepstral coefficients used by the decoder, or for just some of them, typically those of smallest orders.

The corresponding means of the decoder are illustrated by figure 19. The decoder operates essentially like that described with reference to figure 8 to determine the signal frames of integer rank. An interpolation module 124 identical to the module 120 of the coder estimates the intermediate coefficients cx_i[n-1/2] from the quantized coefficients cx_q[n-1] and cx_q[n] provided by the module 47 and/or the module 48 from the indices icxs, icxi extracted from the stream Φ. A module for extracting parameters 125 receives the quantization index icx[n-1/2] from the input demultiplexer 45 of the decoder, and deduces therefrom the quantized interpolation error ecx_q[n-1/2] from the same quantization dictionary as that used by the module

122 of the coder. An adder 126 sums the cepstral vectors cx_i[n-1/2] and ecx_q[n-1/2] so as to provide the cepstral coefficients cx[n-1/2] which will be used by the decoder (modules 51-57, 95, 96, 115 and/or modules 85-87, 92, 95, 96, 115) so as to form the interpolated frame of rank n-1/2.

If just some of the cepstral coefficients have formed the subject of an interpolation error quantization, the others are determined by the decoder by a simple interpolation with no correction.

The decoder can also interpolate the other parameters F_0, Emix used to synthesize the signal frames. The fundamental frequency F_0 can be linearly interpolated, either in the time domain, or (preferably) directly in the frequency domain. For the possible interpolation of the energy weighting vector Emix, it is appropriate to perform the interpolation after denormalization and while of course taking account of the time shifts between frames.

It should be noted that it is especially advantageous, in order to interpolate the representation of the spectral envelopes, to perform this interpolation in the cepstral domain. Unlike an interpolation performed on other parameters, such as the LSP coefficients (standing for "Line Spectrum Pairs"), the linear interpolation of the cepstral coefficients corresponds to the linear interpolation of the compressed spectral amplitudes.

In the variant represented in figure 20, the coder uses the cepstral vectors cx_q[n], cx_q[n-1], ..., cx_q[n-r] and cx_q[n-1/2] calculated for the last frames which have passed ($r \geq 1$) so as to identify an optimal interpolator filter which, when fed with the quantized cepstral vectors cx_q[n-r], ..., cx_q[n] relating to the frames of integer rank, delivers an interpolated

cepstral vector cx_i[n-1/2] which exhibits a minimum distance with the vector cx[n-1/2] calculated for the last frame of half-integer rank.

In the example represented in figure 20, this interpolator filter 128 is present in the coder, and a subtractor 129 deducts its output cx_i[n-1/2] from the calculated cepstral vector cx[n-1/2]. A minimization module 130 determines the parameter set {P} of the interpolator filter 128, for which the interpolation error ecx[n-1/2] delivered by the subtractor 129 exhibits a minimum norm. This parameter set {P} is addressed to a quantization module 131 which provides a corresponding quantization index iP to the output multiplexer 6 of the coder.

As a function of the bit rate allocated in the stream Φ to the indices for quantizing the parameters {P} defining the optimal interpolator filter 128, it will be possible to adopt a finer or coarser quantization of these parameters, or a more or less elaborate form of the interpolator filter, or else to envisage several interpolator filters quantized differently for various vectors of cepstral coefficients.

In a simple embodiment, the interpolator filter 128 is linear, with r = 1:

$$cx\_i[n-1/2] = \rho . cx\_q[n-1] + (1-\rho) . cx\_q[n] \quad (23)$$

and the parameter set {P} is limited to the coefficient ρ lying between 0 and 1.

From the indices iP for quantizing the parameters {P} obtained in the binary stream φ, the decoder reconstructs the interpolator filter 128 (to within quantization errors) and processes the spectral vectors cx_q[n-r], ..., cx_q[n] so as to estimate the cepstral

coefficients cx[n-1/2] used to synthesize the frames of half-integer rank.

Generally, the decoder can use a simple interpolation method (without transmission of parameters by the coder for the frames of half-integer rank), and an interpolation method with incorporation of a quantized interpolation error (according to figures 17 and 18), or an interpolation method with an optimal interpolator filter (according to figure 19) to evaluate the frames of half-integer rank in addition to the frames of integer rank evaluated directly, as explained with reference to figures 8 to 13. The temporal synthesis module 116 can then combine the collection of these frames evaluated so as to form the synthesized signal $\hat{x}$ in the manner explained hereinbelow with reference to figures 14, 21 and 22.

As in the method of temporal synthesis described above, the module 116 performs an overlap sum of frames modified with respect to those evaluated successively at the output of the module 115, and this modification can be viewed in two steps of which the first is identical to that described above with reference to figure 14 (divide the samples of the frame 2′ by the analysis window f_A).

The second step (figure 21) consists in multiplying the samples of the renormalized frame 2″ by a synthesis window $f_S^{'}$ satisfying the following properties:

$$f_S^{'}(i) = 0 \text{ for } 0 \le i < N/2 - M/p \text{ and } N/2 + M/p \le i < N \qquad (24)$$

$$f_S^{'}(i) + f_S^{'}(i + M/p) = A \text{ for } N/2 - M/p \le i < N/2 \qquad (25)$$

where A designates an arbitrary positive constant, for example A = 1 and p is the integer such that the time

shift between the successive frames (calculated directly and interpolated) is M/p samples, i.e. p = 2 in the example described. The synthesis window $f_s^{'}(i)$ increases progressively for i going from N/2 - M/p to N/2. It is, for example, a raised sinusoid on the interval $N/2 - M/p \leq i < N/2 + M/p$. In particular, the synthesis window $f_s^{'}$ can, over this interval, be a Hamming window (as represented in figure 21) or a Hanning window.

Figure 21 shows the successive frames 2″ repositioned over time by the module 116. The hatching indicates the removed portions of the frames (synthesis window at 0). It may be seen that by performing the overlap sum of the samples of the successive frames, the property (25) ensures homogeneous weighting of the samples of the synthesized signal.

As in the method of synthesis illustrated by figures 14 and 15, the procedure for weighting the frames obtained by inverse Fourier transform of the spectra Y can be performed in a single step, with a compound window $f_C^{'}(i) = f_S^{'}(i)/f_A(i)$. Figure 22 shows the form of the compound window $f_C^{'}$ in the case where the windows f_A and $f_S^{'}$ are of Hamming type.

Like the method of temporal synthesis illustrated by figures 14 to 17, that illustrated by figures 14, 21 and 22 makes it possible to take into account an overlap L between two analysis frames (for which the analysis is performed completely) which is smaller than half the size N of these frames. In general, this latter method is applicable when the successive analysis frames exhibit mutual time shifts M of more than N/2 samples (possibly even of more than N samples if a very low bit rate is required), the interpolation leading to a collection of frames whose mutual time shifts are less than N/2 samples.

The interpolated frames can form the subject of a reduced transmission of coding parameters, as is described above, but this is not compulsory. This embodiment makes it possible to retain a relatively large interval M between two analysis frames, and hence to limit the transmission bit rate required, whilst limiting the discontinuities which are liable to appear by virtue of the size of this interval with respect to the typical timescales for the variations in the parameters of the audio signal, in particular the cepstral coefficients and the fundamental frequency.

CLAIMS

1. A method of analyzing an audio signal (x) processed by successive frames of N samples, in which the samples of each frame are weighted by an analysis window (f_A) of Hamming, Hanning, Kaiser or similar type, a spectrum of the audio signal is calculated by transforming each frame of weighted samples in the frequency domain, and the spectrum of the audio signal is processed so as to deliver parameters (cx_sup, cx_inf, Emix) for synthesizing a signal derived from the analyzed audio signal, characterized in that the successive frames comprise an alternation of frames for which are delivered complete sets of synthesis parameters and of frames for which are delivered incomplete sets of synthesis parameters, and in that the successive frames for which complete sets of synthesis parameters are delivered exhibit mutual overlaps of less than N/2 samples.

2. The method as claimed in claim 1, in which the incomplete sets of synthesis parameters include data (icx[n-1/2]) representing an error (ecx[n-1/2]) of interpolation of at least one of the synthesis parameters.

3. The method as claimed in claim 1, in which the incomplete sets of synthesis parameters include data (iP) representing a filter (128) for interpolating at least one of the synthesis parameters.

4. The method as claimed in any one of claims 1 to 3, in which the processing of the spectrum of the audio signal (x) comprises an extraction of coding parameters (cx_sup, cx_inf, Emix) with a view to the transmission and/or the storage of the coded audio signal.

5. The method as claimed in any one of claims 1 to 3, in which the processing of the spectrum of the audio signal (x) comprises a denoising by spectral subtraction.

6. An audio processing device, comprising analysis means for executing a method as claimed in any one of claims 1 to 5.

7. A method of synthesizing an audio signal, in which successive spectral estimates (Y) corresponding respectively to frames of N samples of the audio signal which are weighted by an analysis window (f_A) are obtained, the successive frames exhibiting mutual overlaps of L samples, each frame of the audio signal is evaluated by transforming the spectral estimates in the time domain, and the frames evaluated are combined to form the synthesized signal ($\hat{x}$), characterized in that each evaluated frame is modified by applying thereto a processing corresponding to a division by said analysis window (f_A) and to a multiplication by a synthesis window (f_S), and the synthesized signal is formed as an overlap sum of the modified frames, and in that, the number L being smaller than N/2 and the samples of a frame having ranks i numbered from 0 to N-1, the synthesis window $f_S(i)$ satisfies $f_S(N-L+i) + f_S(i) = A$ for $0 \leq i < L$, and is equal to A for $L \leq i < N-L$, A being a positive constant.

8. The method as claimed in claim 7, in which the synthesis window $f_S(i)$ increases from 0 to A for i going from 0 to L.

9. The method as claimed in claim 8, in which the synthesis window $f_S(i)$ for $0 \leq i < L$ is a raised half-sinusoid.

10. A method of synthesizing an audio signal, in which a set of successive overlapping frames of N samples of the audio signal which are weighted by an analysis window (f_A) is evaluated, by transforming in the time domain spectral estimates (Y) corresponding respectively to said frames, and the evaluated frames are combined to form the synthesized signal ($\hat{x}$), characterized in that, for a subset of the evaluated frames, the spectral estimates are obtained by processing synthesis parameters (cx_sup_q, cx_inf_q, Emix) respectively associated with the frames of said subset while, for the frames which do not form part of the subset, the spectral estimates are obtained with an interpolation of a part at least of the synthesis parameters, in that the successive frames of said subset exhibit mutual time shifts of M samples, the number M being larger than N/2, while the successive frames of said set exhibit mutual time shifts of M/p samples, p being an integer larger than 1, in that each evaluated frame is modified by applying thereto a processing corresponding to a division by said analysis window (f_A) and to a multiplication by a synthesis window (f'_S), and the synthesized signal is formed as an overlap sum of the modified frames, and in that, the samples of a frame having ranks i numbered from 0 to N-1, the synthesis window $f'_S(i)$ has a support limited to the ranks i ranging from N/2 - M/p to N/2 + M/p and satisfies $f'_S(i) + f'_S(i + M/p) = A$ for $N/2 - M/p \le i < N/2$, A being a positive constant.

11. The method as claimed in claim 10, in which the synthesis window $f'_S(i)$ increases for i ranging from N/2 - M/p to N/2.

12. The method as claimed in claim 11, in which the synthesis window $f'_s(i)$ for $N/2 - M/p \le i < N/2 + M/p$ is a raised sinusoid.

13. The method as claimed in any one of claims 10 to 12, in which data (icx_q[n-1/2]) representing an interpolation error (ecx_q[n-1/2]) are associated with the frames which do not form part of said subset, and are used to correct at least one of the interpolated synthesis parameters (cx_i[n-1/2]).

14. The method as claimed in any one of claims 10 to 12, in which data (iP) representing an interpolator filter (128) are associated with the frames which do not form part of said subset, and are used to interpolate at least one of the synthesis parameters.

15. The method as claimed in any one of claims 10 to 14, in which the synthesis parameters comprise cepstral coefficients (cx[n]) subjected to the interpolation.

16. An audio processing device, comprising synthesis means for executing a method as claimed in any one of claims 7 to 15.

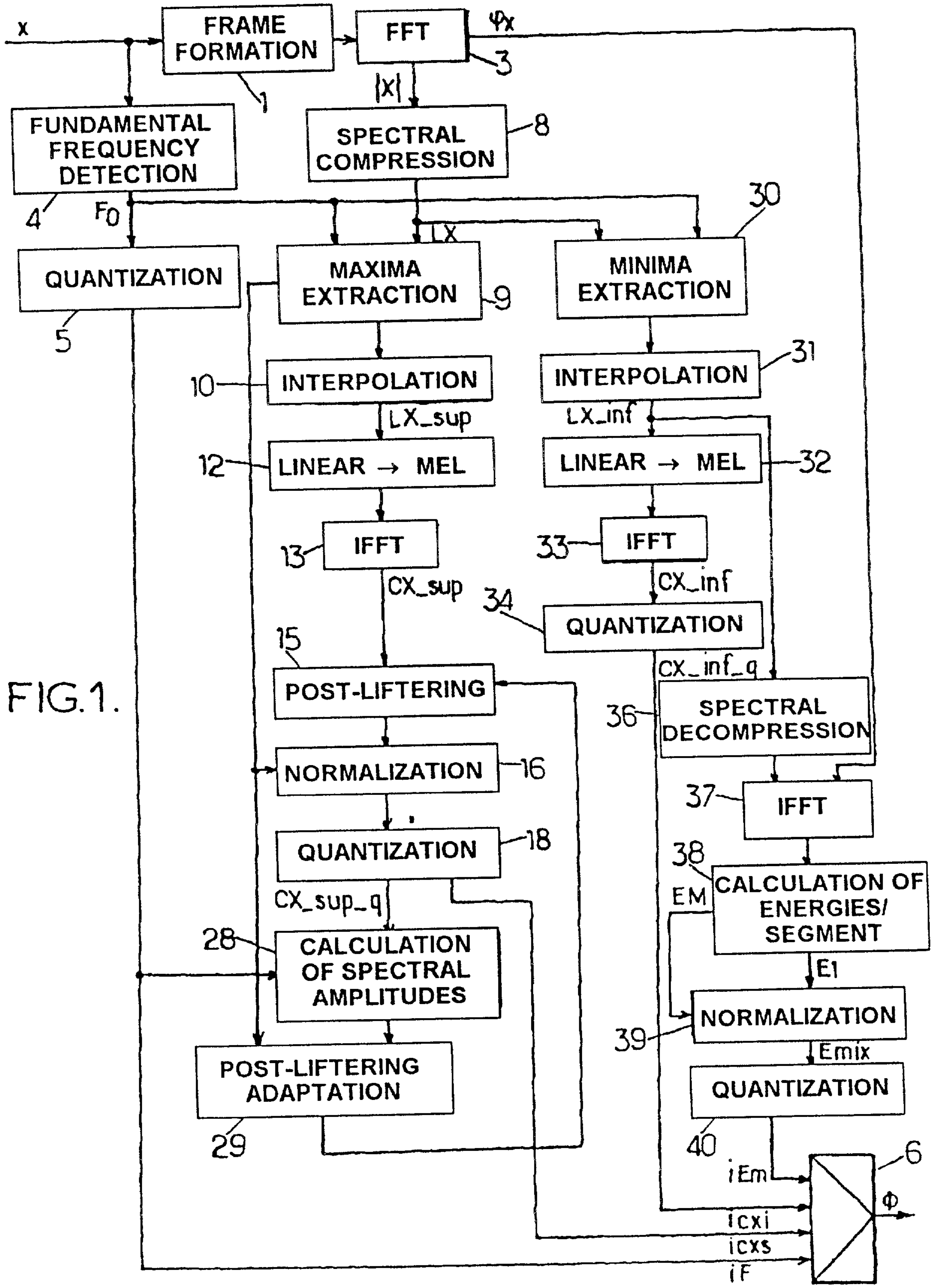

x
FRAME FORMATION
FFT
φx
3
1
|X|
FUNDAMENTAL FREQUENCY DETECTION
SPECTRAL COMPRESSION
8
4
F0
30
LX
QUANTIZATION
MAXIMA EXTRACTION
MINIMA EXTRACTION
9
5
INTERPOLATION
INTERPOLATION
10
31
LX_sup
LX_inf
LINEAR → MEL
LINEAR → MEL
12
32
IFFT
IFFT
13
33
CX_sup
CX_inf
34
QUANTIZATION
15
CX_inf_q
POST-LIFTERING
SPECTRAL DECOMPRESSION
36
NORMALIZATION
16
37
IFFT
QUANTIZATION
18
38
CALCULATION OF ENERGIES/ SEGMENT
EM
CX_sup_q
28
CALCULATION OF SPECTRAL AMPLITUDES
E1
NORMALIZATION
39
Emix
POST-LIFTERING ADAPTATION
QUANTIZATION
29
40
6
iEm
Φ
icxi
icxs
iF

FIG.1.

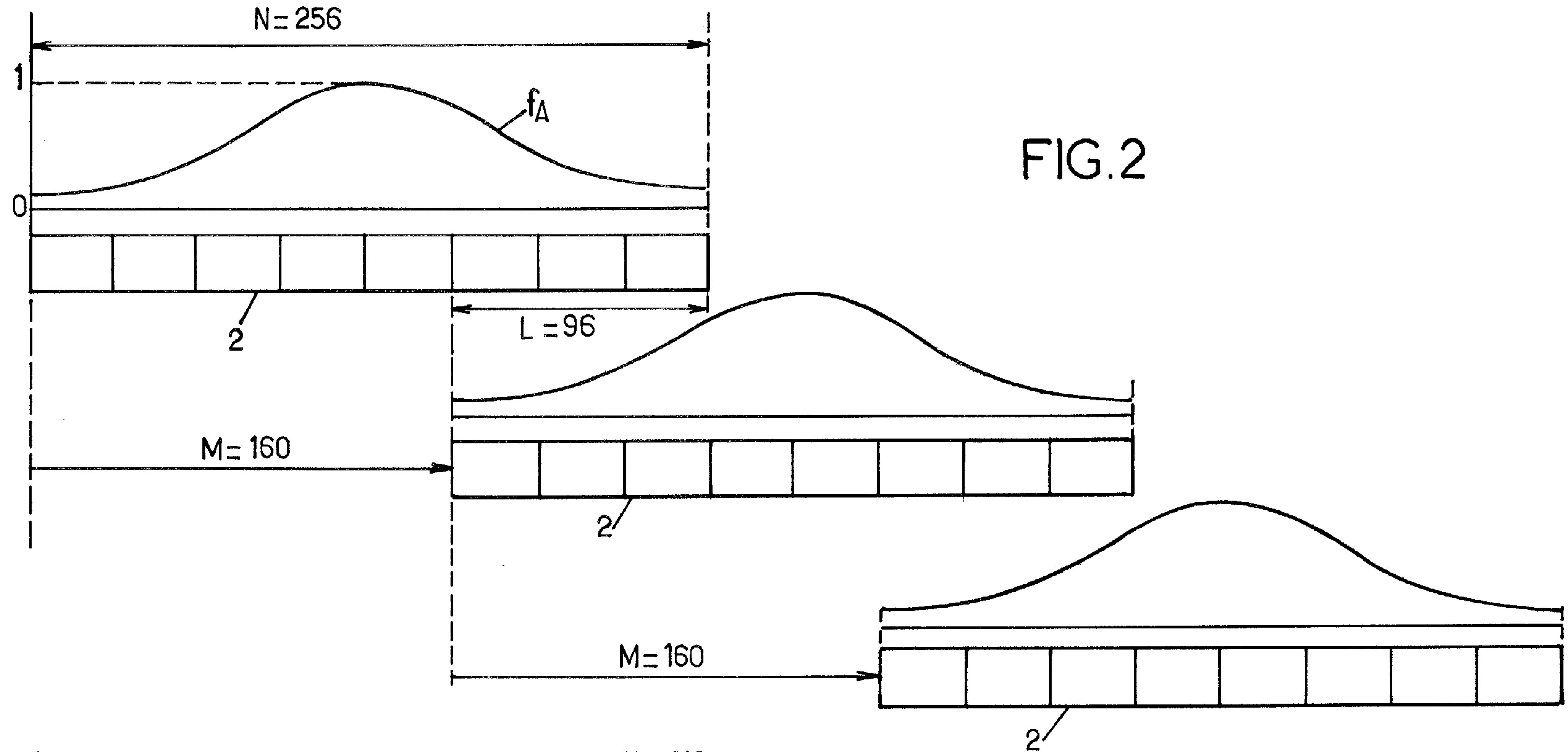


FIG.2

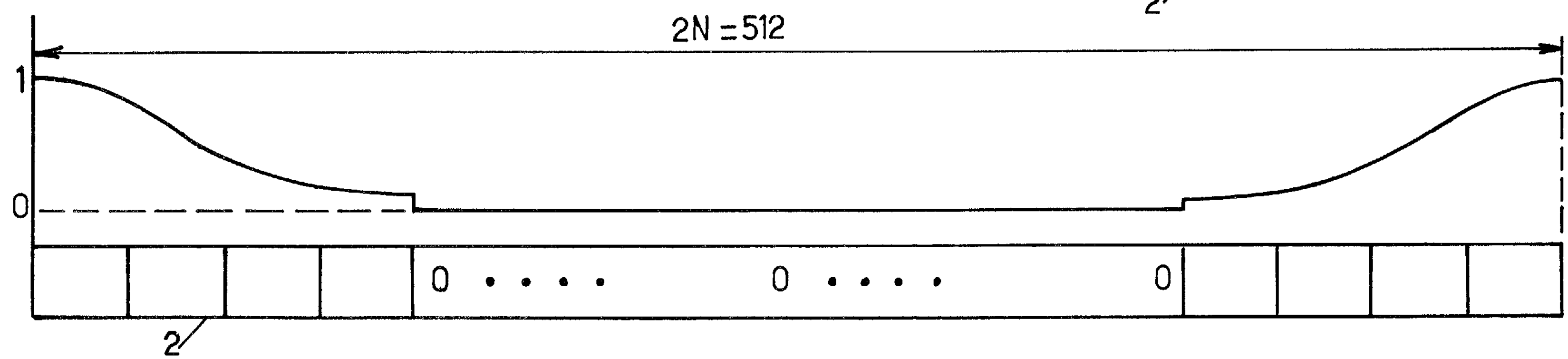


FIG.3.

FIG.4.

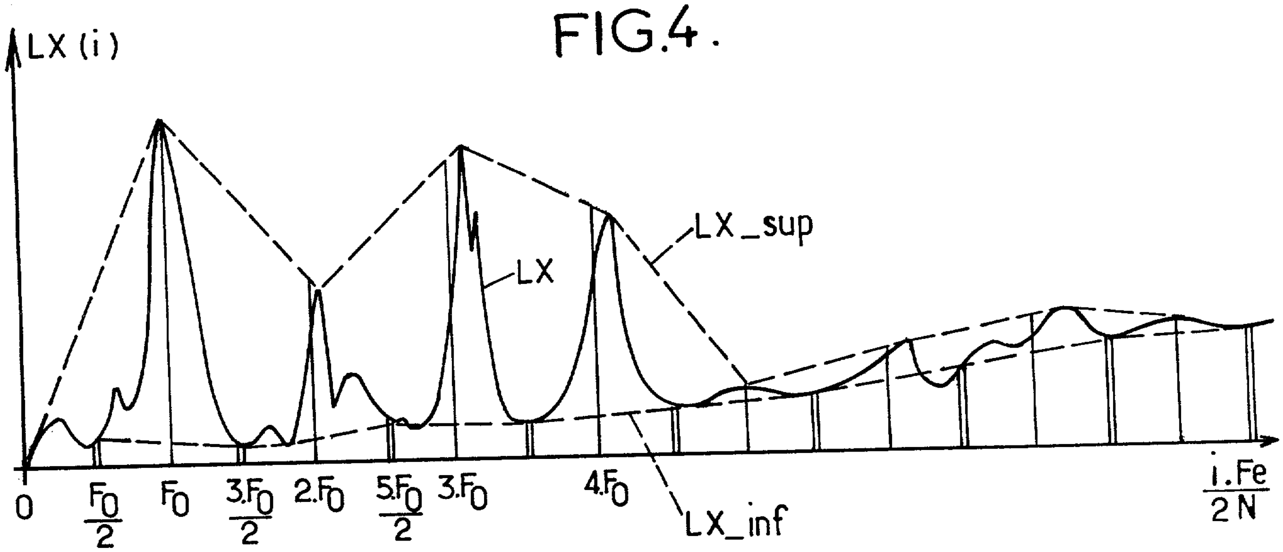
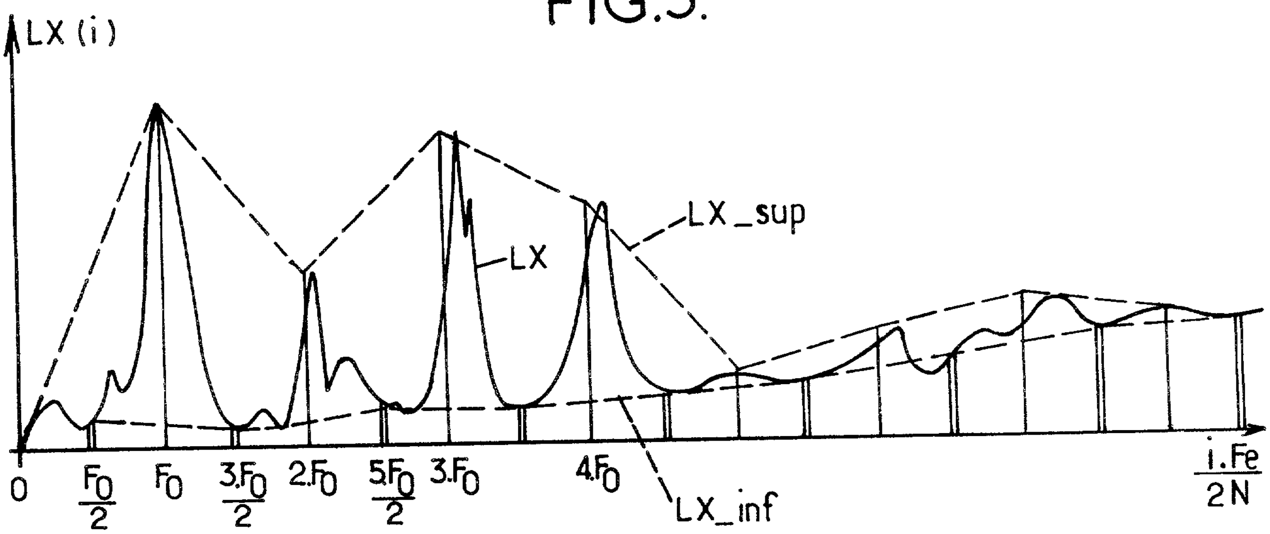
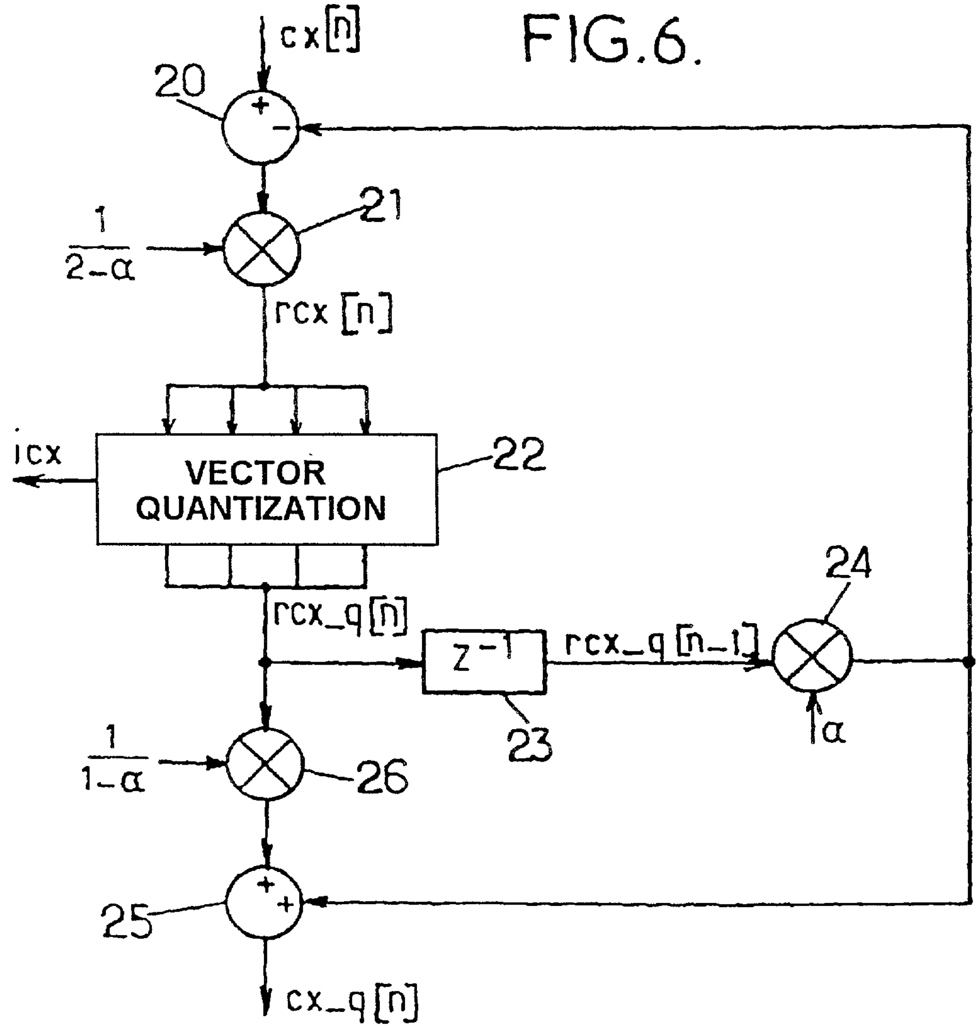

LX (i)
LX_sup
LX
LX_inf
0
$\frac{F_0}{2}$
F_0
$\frac{3.F_0}{2}$
$2.F_0$
$\frac{5.F_0}{2}$
$3.F_0$
$4.F_0$
$\frac{i.Fe}{2N}$

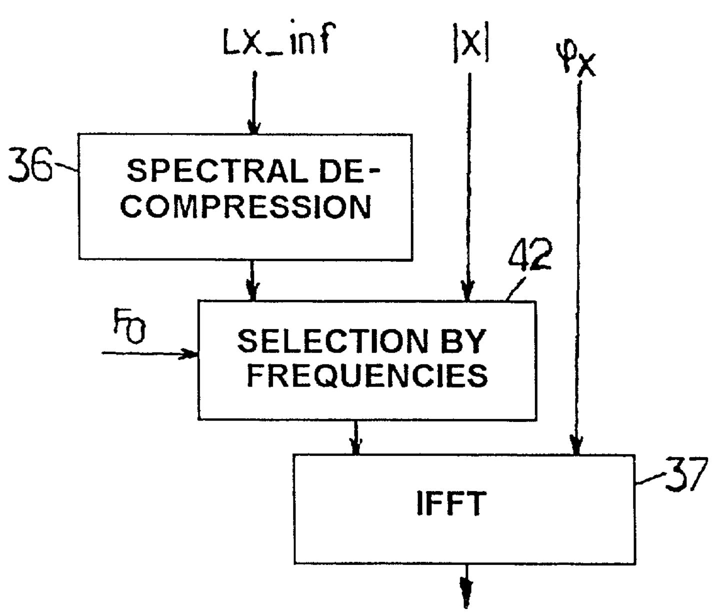
FIG.5.


LX (i)
LX_sup
LX
LX_inf
0
$\frac{F_0}{2}$
F_0
$\frac{3.F_0}{2}$
$2.F_0$
$\frac{5.F_0}{2}$
$3.F_0$
$4.F_0$
$\frac{i.Fe}{2N}$



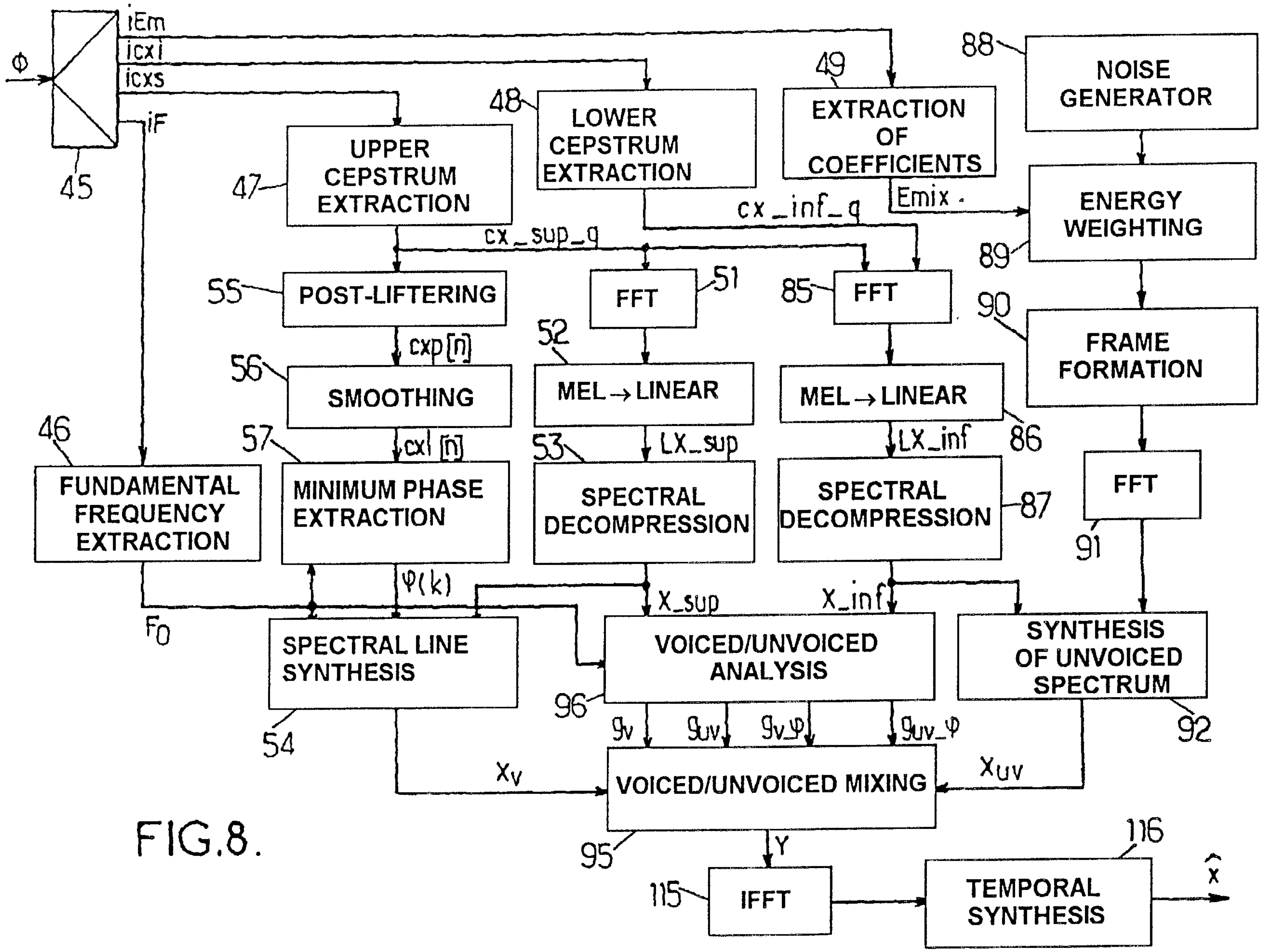
cx[n]
20
21
$\frac{1}{2-\alpha}$
rcx [n]
icx
VECTOR QUANTIZATION
22
rcx_q[n]
z^-1
rcx_q[n-1]
24
α
23
$\frac{1}{1-\alpha}$
26
25
cx_q[n]

FIG.6.



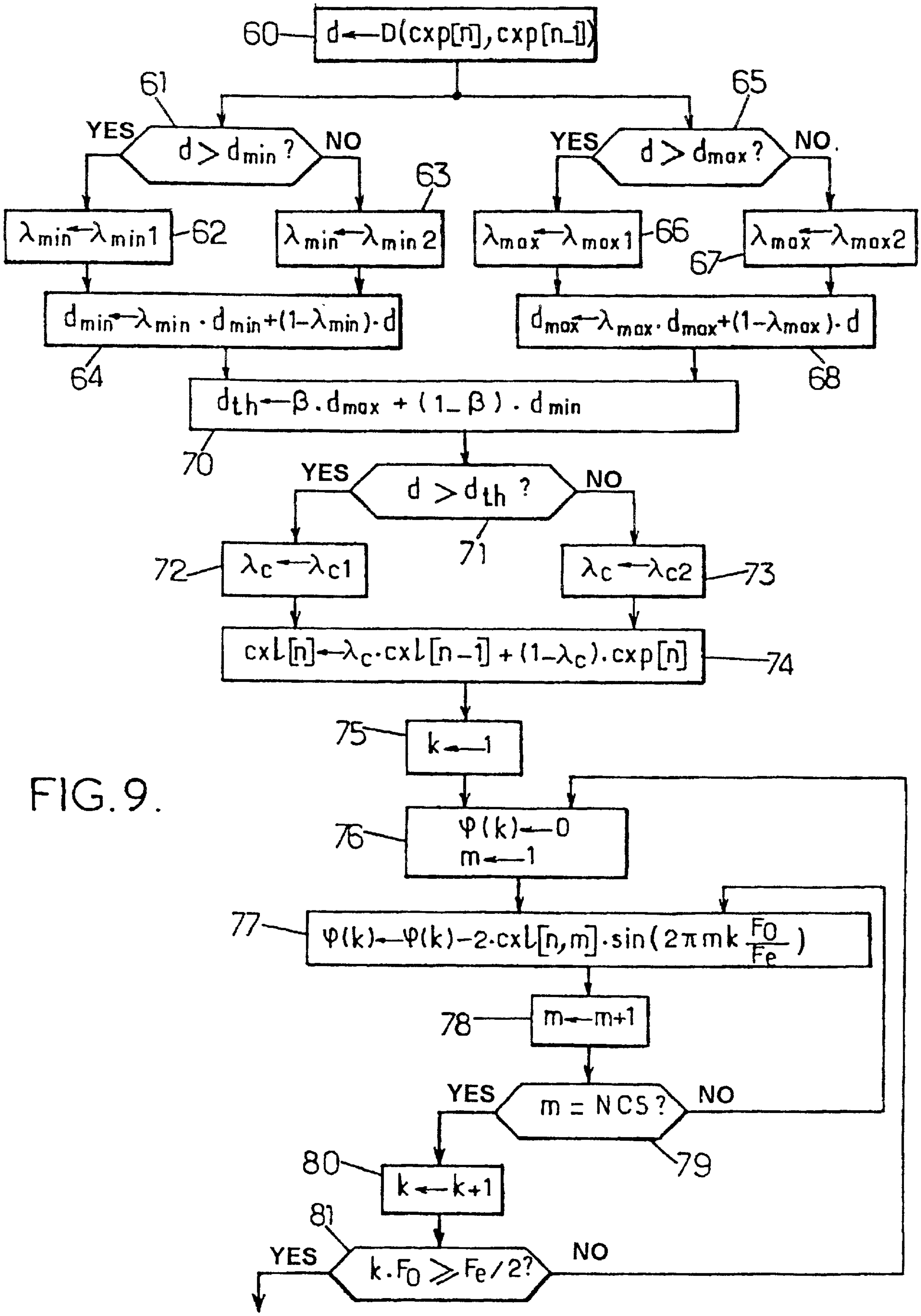
Lx_inf
|X|
φx
36
SPECTRAL DE-COMPRESSION
42
F0
SELECTION BY FREQUENCIES
IFFT
37

FIG.7.



φ
iEm
icxi
icxs
iF
45
UPPER CEPSTRUM EXTRACTION
47
LOWER CEPSTRUM EXTRACTION
48
EXTRACTION OF COEFFICIENTS
49
NOISE GENERATOR
88
Emix
ENERGY WEIGHTING
89
cx_inf_q
cx_sup_q
POST-LIFTERING
55
FFT
51
FFT
85
FRAME FORMATION
90
cxp[n]
SMOOTHING
56
MEL→LINEAR
52
MEL→LINEAR
86
46
57
cxl[n]
LX_sup
LX_inf
FFT
91
FUNDAMENTAL FREQUENCY EXTRACTION
MINIMUM PHASE EXTRACTION
SPECTRAL DECOMPRESSION
53
SPECTRAL DECOMPRESSION
87
φ(k)
X_sup
X_inf
F0
SPECTRAL LINE SYNTHESIS
54
VOICED/UNVOICED ANALYSIS
96
SYNTHESIS OF UNVOICED SPECTRUM
92
gv
guv
gv_φ
guv_φ
Xv
Xuv
VOICED/UNVOICED MIXING
95
Y
IFFT
115
TEMPORAL SYNTHESIS
116
x̂

FIG.8.



60
d ← D(cxp[n], cxp[n-1])
61
YES
d > dmin ?
NO
63
65
YES
d > dmax ?
NO
λmin ← λmin1
62
λmin ← λmin2
λmax ← λmax1
66
67
λmax ← λmax2
dmin ← λmin · dmin + (1−λmin) · d
dmax ← λmax · dmax + (1−λmax) · d
64
68
dth ← β . dmax + (1 − β) . dmin
70
YES
d > dth ?
NO
71
72
λc ← λc1
λc ← λc2
73
cxL[n] ← λc · cxL[n−1] + (1−λc) . cxp[n]
74
75
k ← 1
76
φ(k) ← 0
m ← 1
77
φ(k) ← φ(k) − 2 · cxL[n,m] · sin(2πmk F0/Fe)
78
m ← m+1
YES
m = NCS ?
NO
79
80
k ← k+1
81
YES
k . F0 ≥ Fe/2?
NO

FIG. 9.

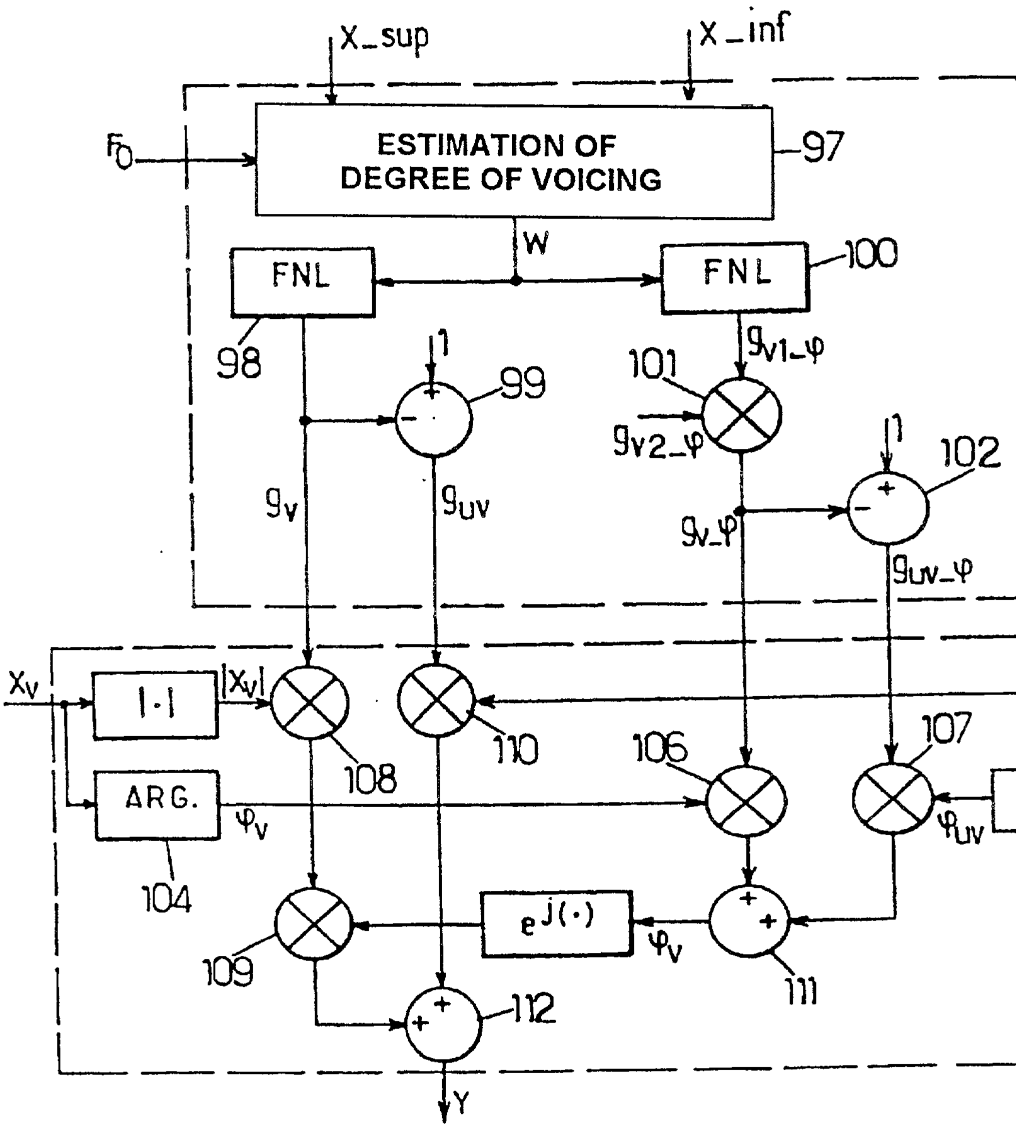


FIG.10.

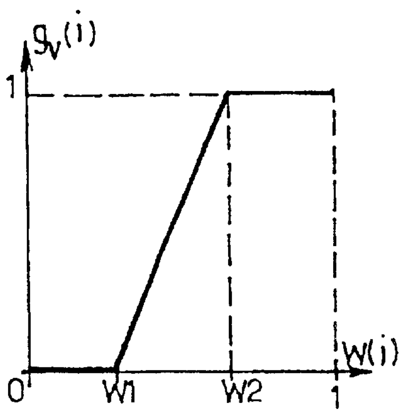


FIG.11

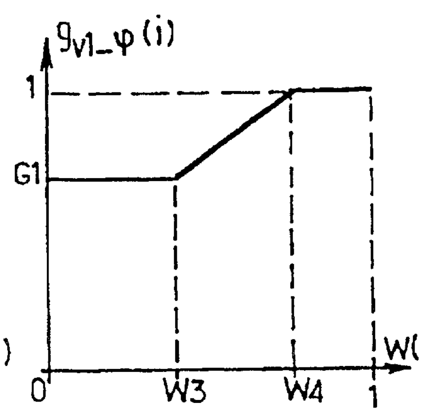


FIG.12.

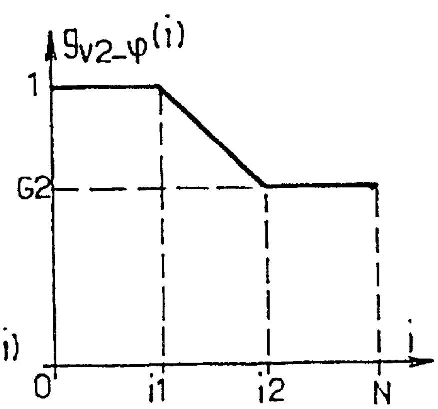


FIG.13.

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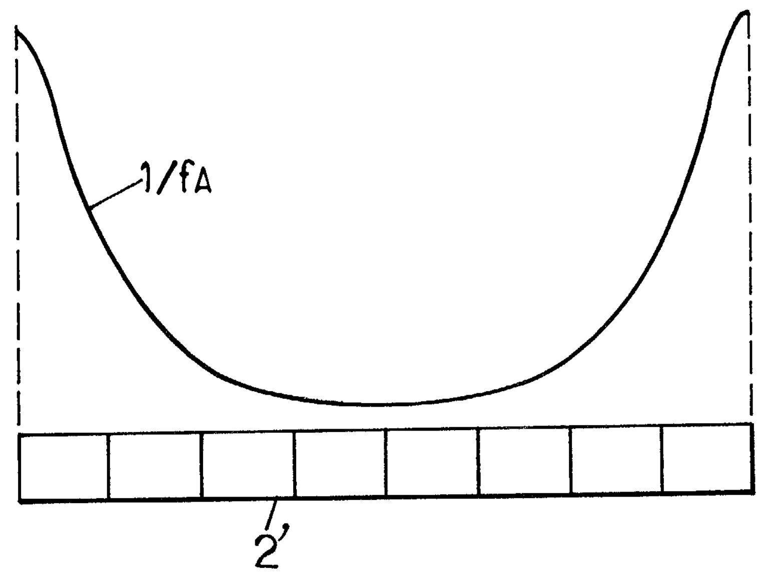

FIG.14.

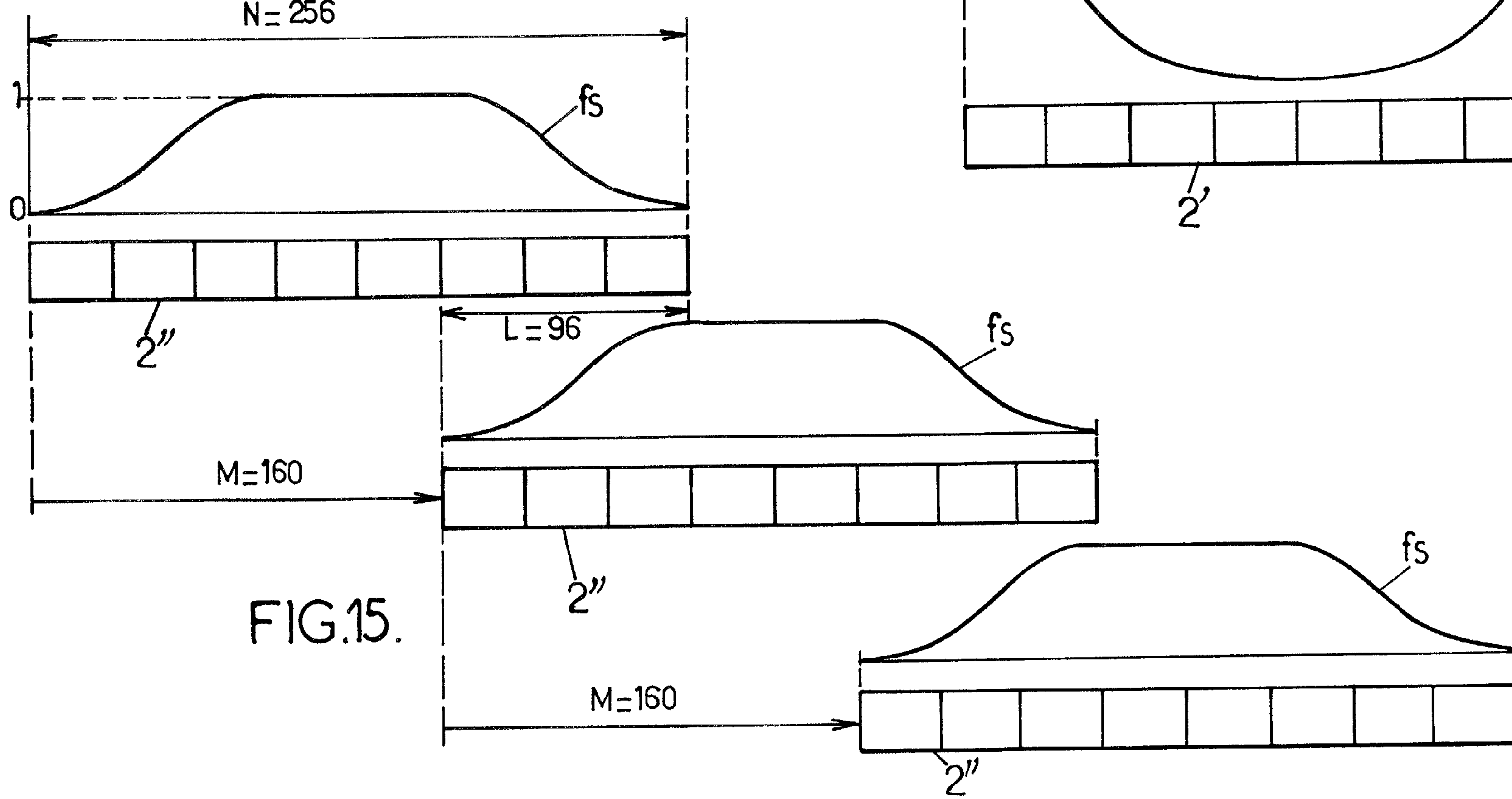

FIG.15.

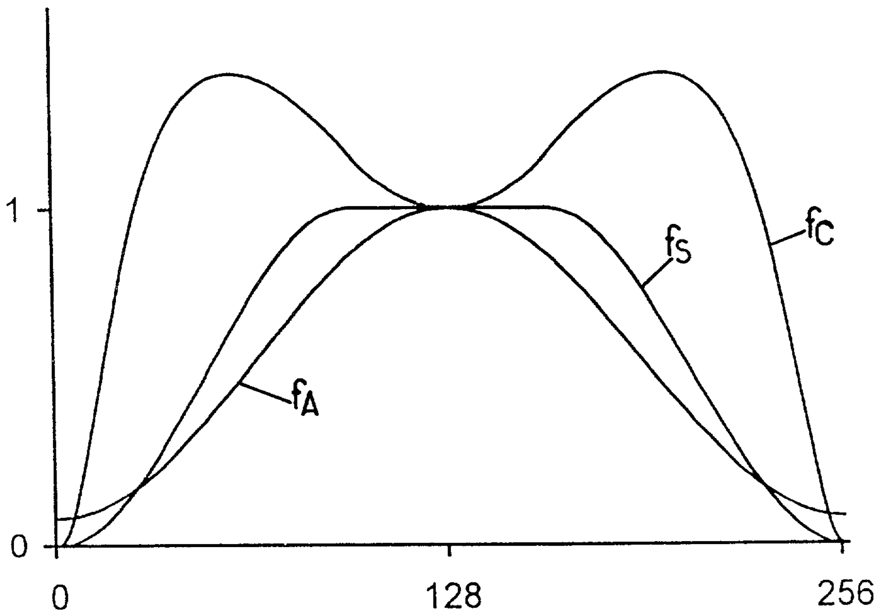

FIG.16.

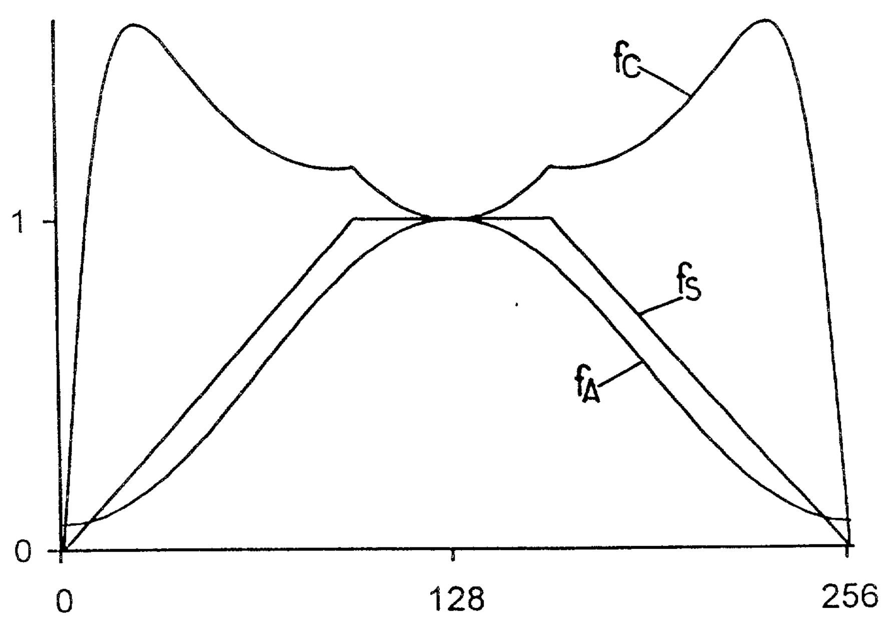

FIG.17.

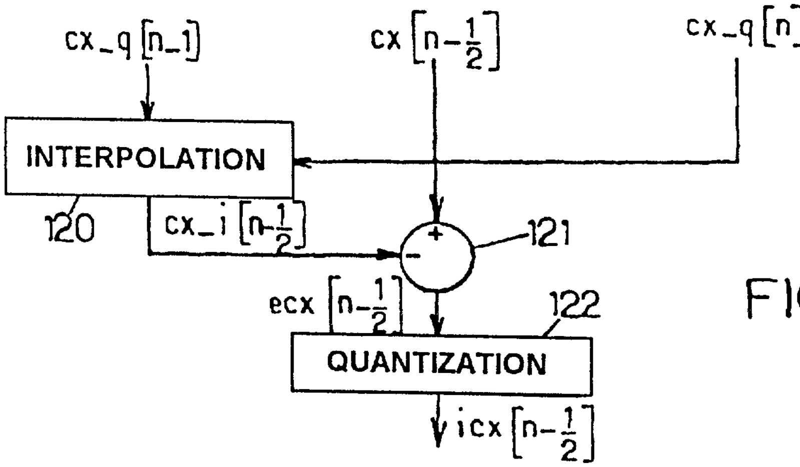

FIG.18.

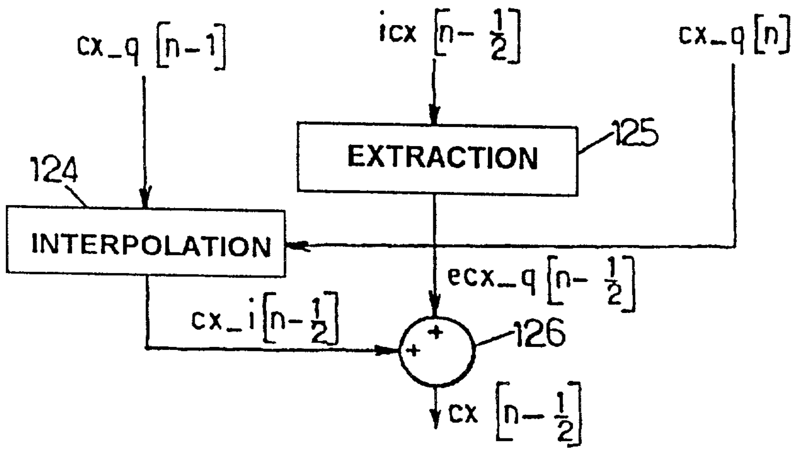

FIG.19.

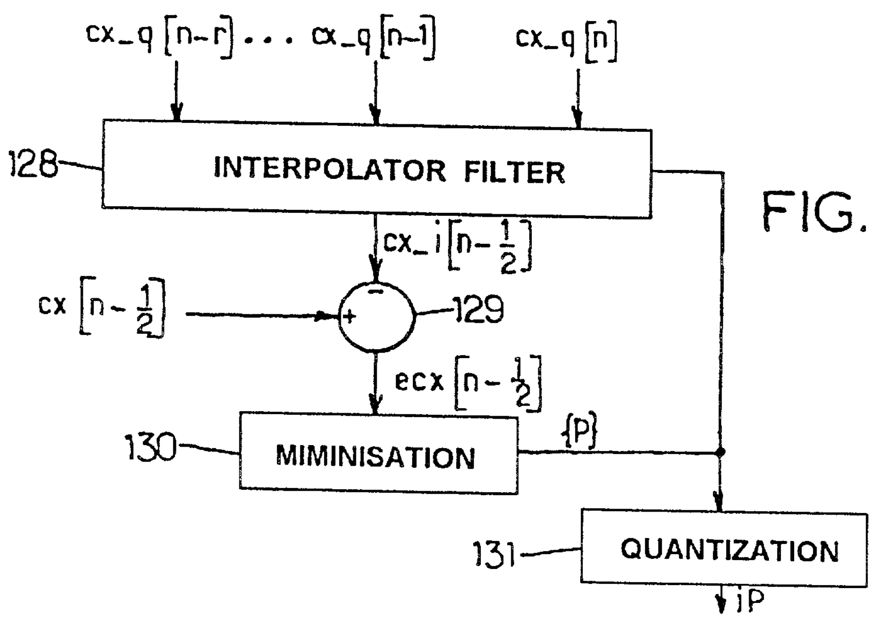

FIG.20.

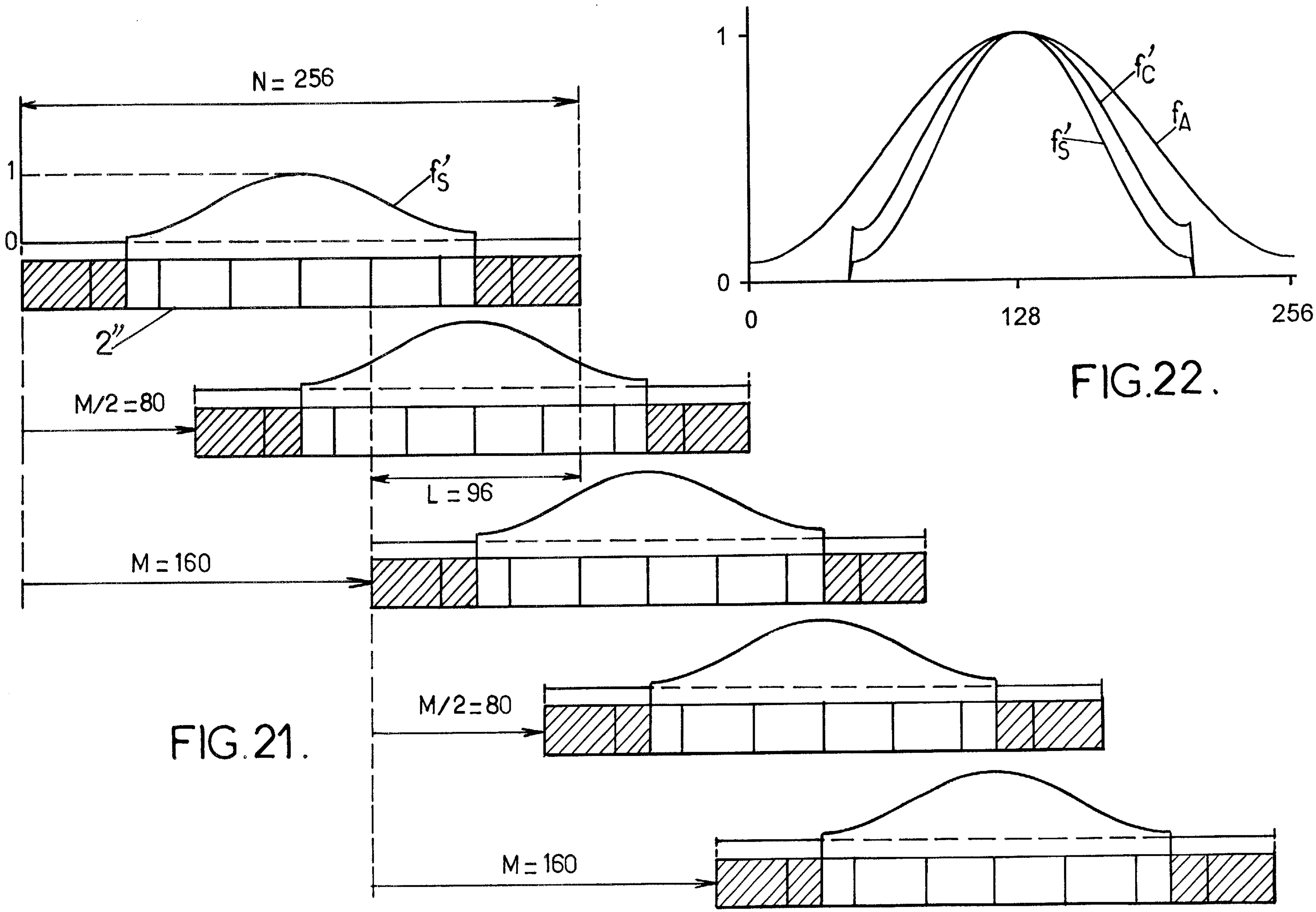

FIG.21.

FIG.22.

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Attorney's Docket No.: US PATENT

DECLARATION AND POWER OF ATTORNEY FOR PATENT APPLICATION

As a below named inventor, I hereby declare that:

My residence, post office address and citizenship are as stated below, next to my name.

I believe I am the original, first, and sole inventor (if only one name is listed below) or an original, first, and joint inventor (if plural names are listed below) of the subject matter which is claimed and for which a patent is sought on the invention entitled
METHODS AND DEVICES FOR AUDIO ANALYSIS AND SYNTHESIS.

the specification of which

X	is attached hereto.
	was filed on ____________ as
	United States Application Number ____________
	Or PCT International Application Number ____________
	And was amended on ____________ (if applicable)

I hereby state that I have reviewed and understand the contents of the above-identified specification, including the claim(s), as amended by any amendment referred to above. I do not know and do not believe that the claimed invention was ever known or used in the United States of America before my invention thereof, or patented or described in any printed publication in any country before my invention thereof or more than one year prior to this application, that the same was not in public use or on sale in the United States of America more than one year prior to this application, and that the invention has not been patented or made the subject of an inventor's certificate Issued before the date of this application in any country foreign to the United States of America on an application filed by me or my legal representatives or assigns more than twelve months (for a utility patent application) or six months (for a design patent application) prior to this application.

I acknowledge the duty to disclose all information known to me to be material to patentability as defined in Title 37, Code of Federal Regulations, Section 1.56.

I hereby claim foreign priority benefits under Title 35, United States Code, Section 119(a)-(d), of any foreign application(s) for patent or inventor's certificate listed below and have also identified below any foreign application for patent or inventor's certificate having a filing date before that of the application on which priority is claimed:

Express Mail No. EL857974173US

Full Name of Sole/First Inventor: **François CAPMAN** 1-OU	
Inventor's Signature: François CAPMAN	Date: 5th December 2001
Residence: **78000 VERSAILLES (France)** FRX	Citizenship: FRENCH
Post Office Address: **47, rue des Etats Généraux– 78000 VERSAILLES (France)**	
Full Name of Second/Joint Inventor: **Carlo MURGIA**	
Inventor's Signature: CARLO MURGIA 2-CU	Date: 5th December 2001
Residence: **IRVINE, CA 92618 (USA)** CA	Citizenship: ITALIAN
Post Office Address: **458 , Saint Vincent – IRVINE CA 92618 (USA)**	
Full Name of Third/Joint Inventor:	
Inventor's Signature:	Date:
Residence:	Citizenship:
Post Office Address:	
Full Name of fourth/Joint Inventor:	
Inventor's Signature:	Date:
Residence:	Citizenship:
Post Office Address:	

Prior Foreign Application(s):			Priority Claimed	
99 08638	FRANCE	5th July 1999	X	
Number	(Country)	(Day/Month/Year Filed)	Yes	No
Number	(Country)	(Day/Month/Year Filed)	Yes	No

I hereby claim the benefit under title 35, United States Code, Section 119(e) of the United States provisional application(s) listed below:

(Application Number)	(Filing Date)
(Application Number)	(Filing Date)

I hereby claim the benefit under Title 35, United States Code, Section 120 of any United States application(s) listed below and, insofar as the subject matter of each of the claims of this application is not disclosed in the prior United States application in the manner provided by the first paragraph of Title 35, United States Code, Section 112, I acknowledge the duty to disclose all information known to me to be material to patentability as defined in Title 37, Code of Federal regulations, Section 1.56 which became available between the filing date of the prior application and the national or PCT International filing date of this application:

FR00/01904	4th July 2000	PENDING
(Application Number)	Filing Date	(Status-patented, pending, abandoned)

I hereby appoint Timothy N. Trop, Reg. No. 28,994; Fred G. Pruner, Jr., Reg. No. 40,779, Dan C. Hu, Reg. No. 40,025 and Ruben S. Bains, Reg. No. 46,532; my patent attorneys, of TROP, PRUNER & HU, P.C., with offices located at 8554 Katy Freeway, Ste. 100, Houston, TX 77024, telephone (713) 468-8880, my patent attorneys; with full power of substitution and revocation, to prosecute this application and to transact all business in the Patent and Trademark Office connected herewith. 4

Send correspondence to Dan C. Hu, TROP, PRUNER & HU, P.C., 8554 Katy Freeway, Ste. 100, Houston, TX 77024 and direct telephone calls to Dan C. Hu, (713) 468-8880.

I hereby declare that all statements made herein of my own knowledge are true and that all statements made on information and belief are believed to be true; and further that these statements were made with the knowledge that willful false statements and the like so made are punishable by fine or imprisonment, or both, under Section 1001 of Title 18 of the United States Code and that such willful false statements may jeopardize the validity of the application or any patent issued thereon.